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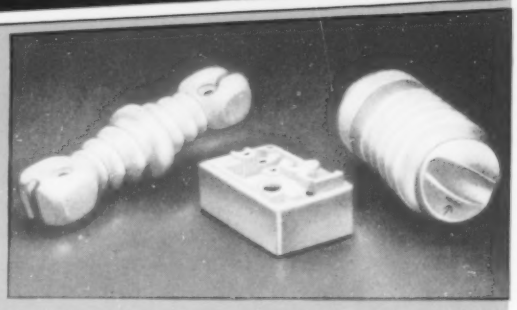

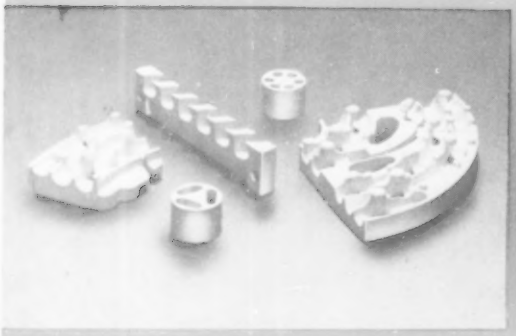

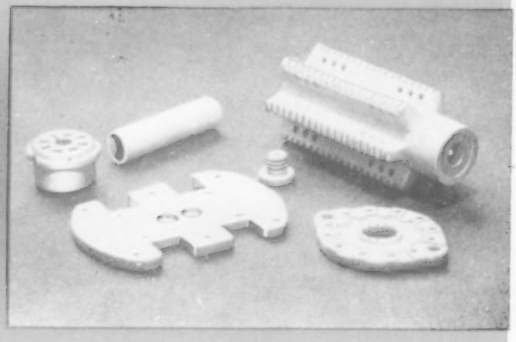

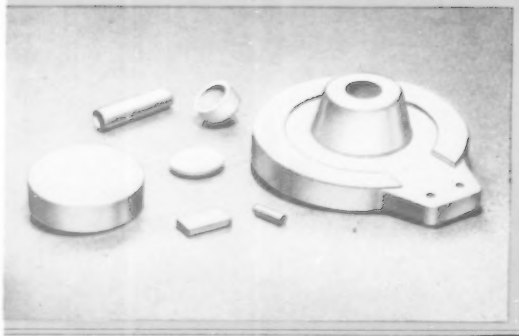

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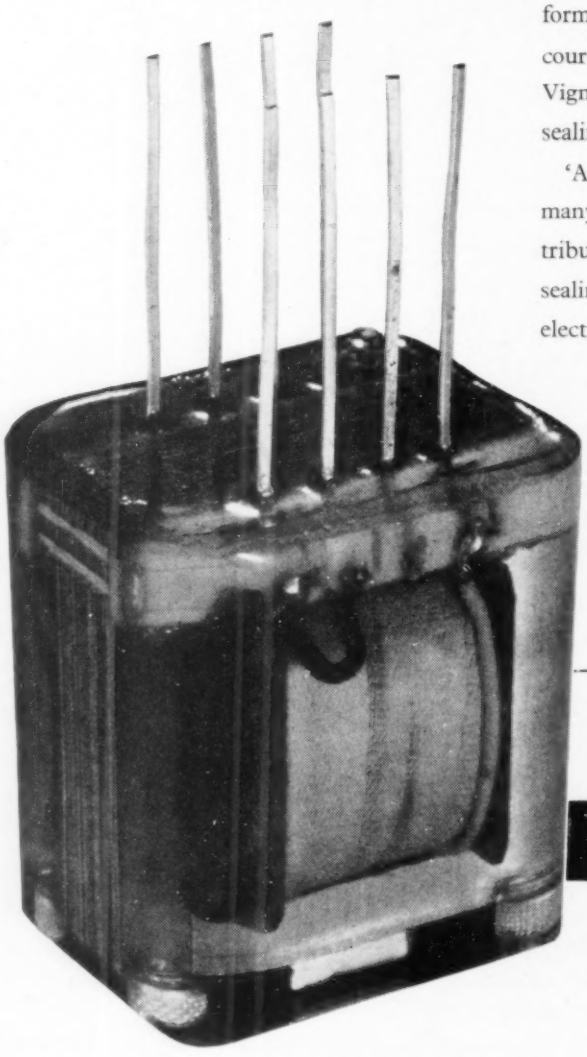

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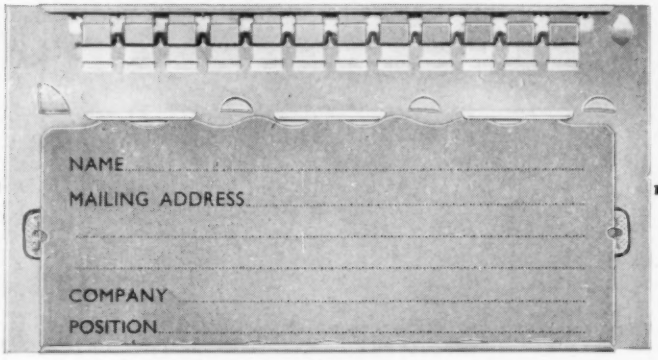

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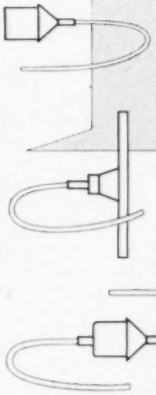

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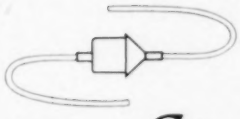

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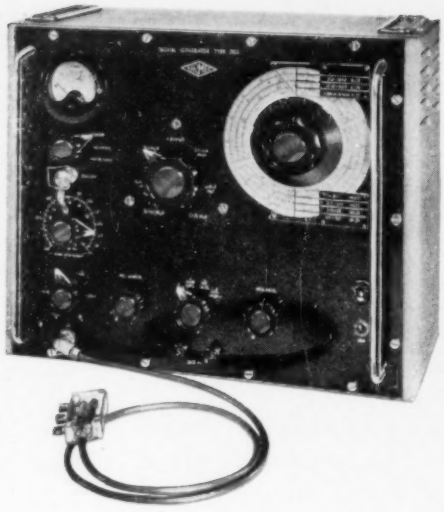

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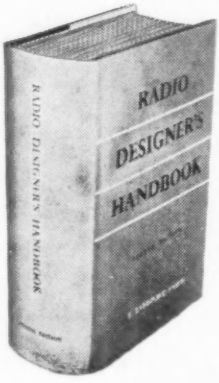

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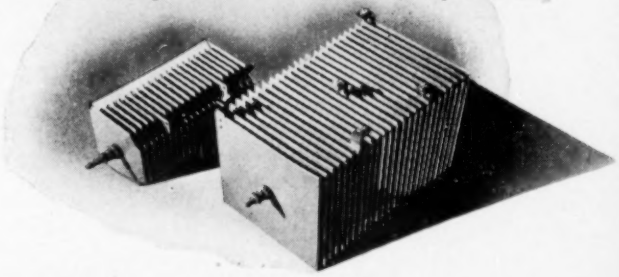

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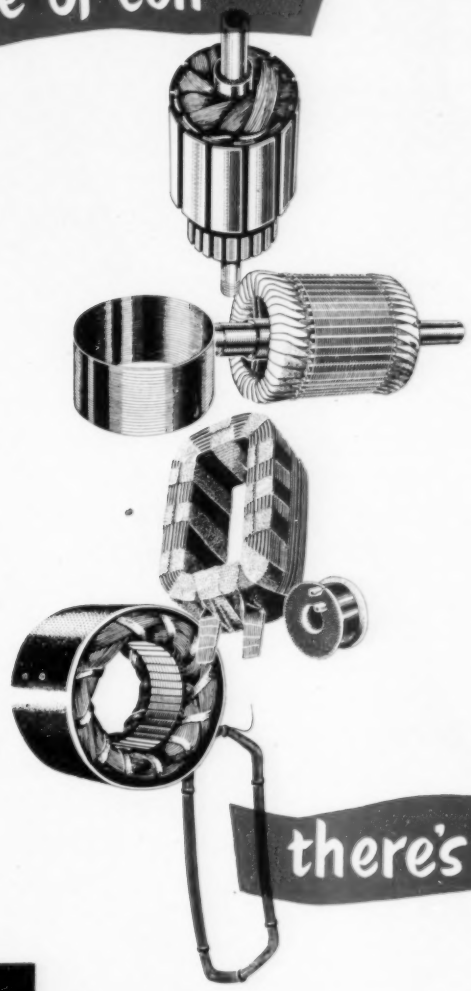

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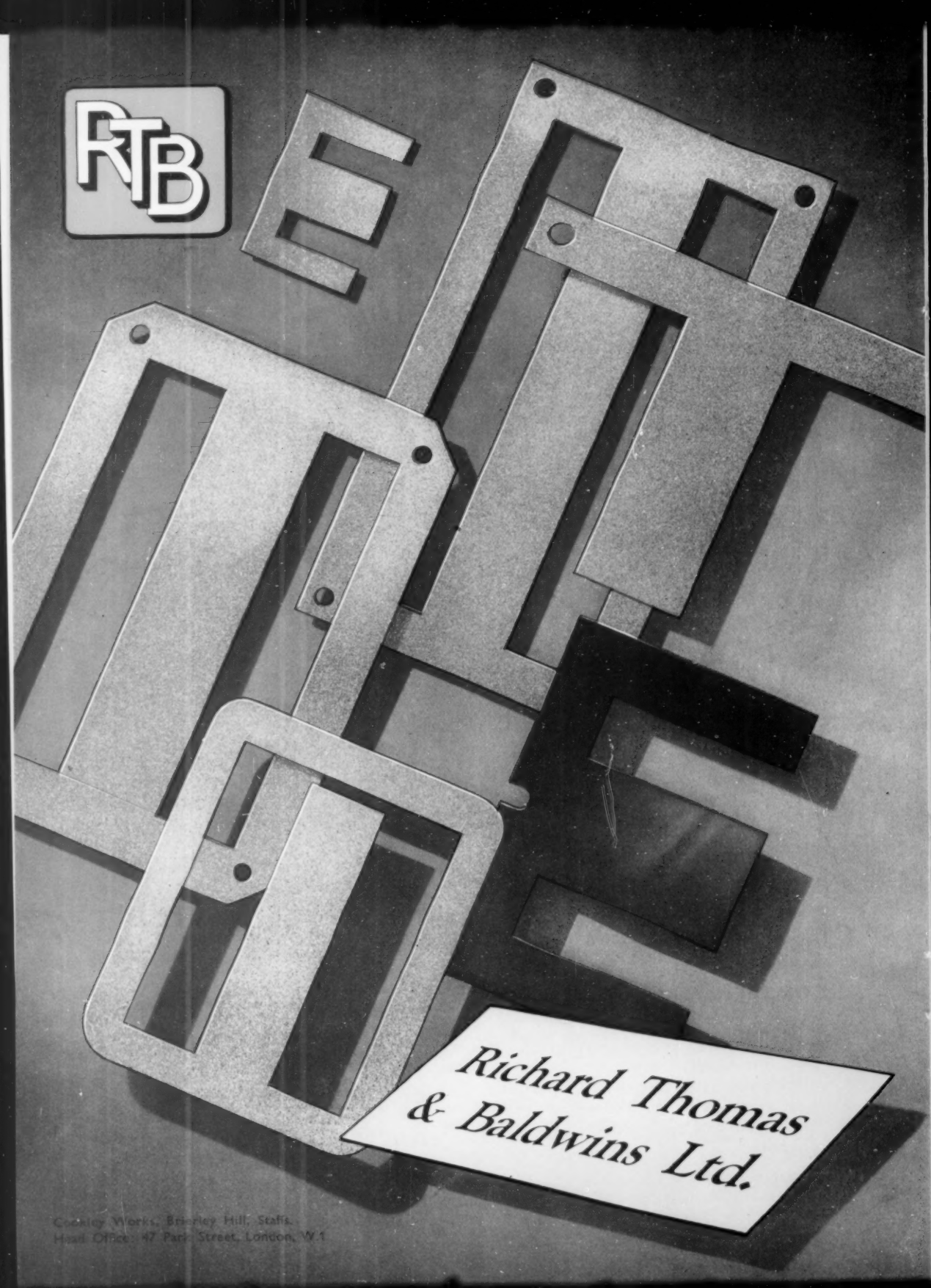

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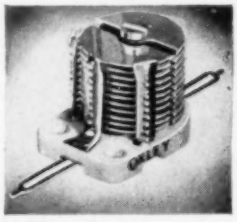

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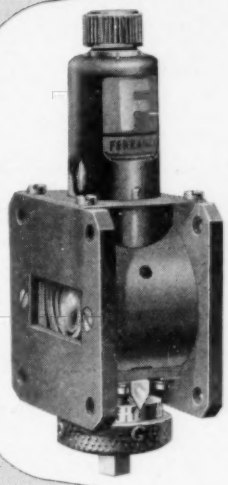

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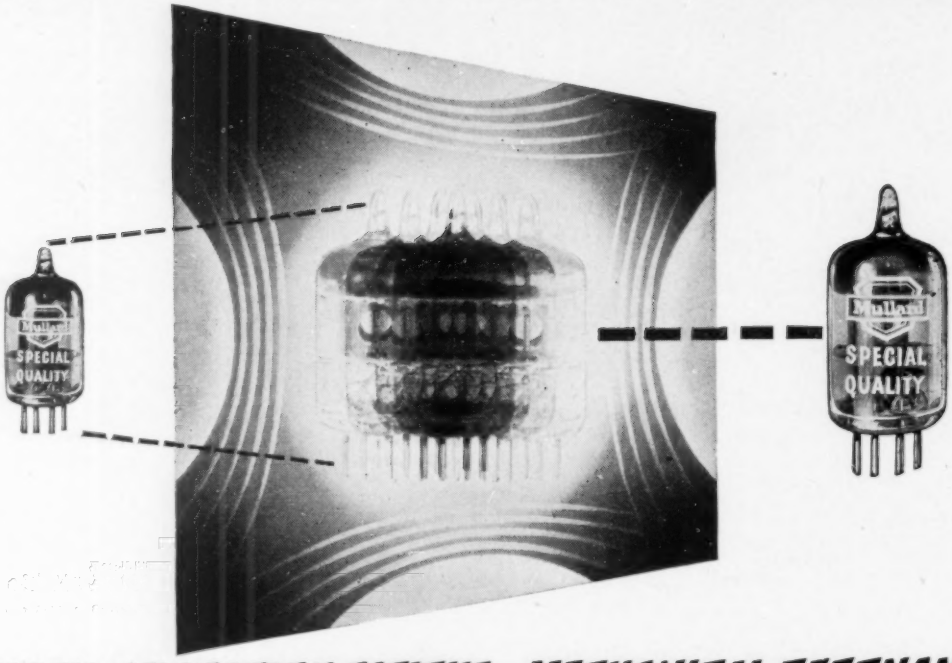

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WIRELESS ENGINEER

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A Heaviside-Lorentz Network Theorem

"THE decomposition of an arbitrary function into the sum of functions of special types has many fascinations. No student of mathematical physics, if he possesses any soul at all, can fail to recognize the poetry that pervades this branch of mathematics." (Oliver Heaviside Electrical Papers, Vol. II, p. 201.)

In our April Editorial we mentioned that in one of his last papers Lorentz developed a generalization of a little-known theorem of Heaviside. At the I.R.E. Convention in Sydney in August 1952, Professor B. D. H. Tellegen of Delft University, read a paper entitled "A General Network Theorem, with Applications", which was published in *Philips Research Reports*, 1952, p. 259, and also in the *Proc. Instn. Radio Engrs.* of Australia, November 1953. This covers a wider field, but one section is devoted to the Heaviside-Lorentz theorem, which we stated as follows:—Given a constant passive network at rest, if at time $t = 0$ a constant e.m.f. is suddenly applied to the network, in general, transients will occur. After a long time t_1 the transients may be considered to have died down, and only a direct current will, in general, be present in the network. The direct current (if there is any) will cause a Joulean heat dissipation at a constant rate per second. If we call W the pseudo heat dissipation which would have occurred if the constant final current had been present all the time from $t = 0$ to $t = t_1$, then the total amount of energy A supplied exceeds W by twice the excess of the stored electric energy U over the stored magnetic energy T; that is $A - W = 2(U - T)$.

We now propose to illustrate this by a simple example. We shall consider first the circuit of Fig. 1, then the circuit of Fig. 2, and finally the combination, as in Fig. 3.

In Fig. 1, on closing the switch, $V - v = R_1 i + L_1\, di/dt$, putting $i = C\, dv/dt$, $V - v = R_1 C\, dv/dt + L_1 C\, d^2v/dt^2$, and

$$\frac{d^2v}{dt^2} + \frac{R_1}{L_1}\frac{dv}{dt} - \frac{V - v}{L_1 C} = 0$$

The solution of this is $V - v = Me^{m_1 t} + Ne^{m_2 t}$. Since, when $t = 0$, $v = 0$ and $dv/dt = i = 0$

$$M + N = V \text{ and } Mm_1 + Nm_2 = 0$$

$$M = V\frac{m_2}{m_2 - m_1};\quad N = V\frac{m_1}{m_1 - m_2}$$

$$m_1 = -\alpha + j\omega;\ m_2 = -\alpha - j\omega$$

where $$\alpha = \frac{R_1}{2L_1};\quad \omega^2 = \frac{1}{L_1 C} - \left(\frac{R_1}{2L_1}\right)^2;$$

$$\alpha^2 + \omega^2 = \frac{1}{L_1 C}.$$

At any moment

$$v = V\left(1 + \frac{m_2}{m_1 - m_2}e^{m_1 t} - \frac{m_1}{m_1 - m_2}e^{m_2 t}\right)$$

and

$$i = C\frac{dv}{dt} = CV\left(\frac{m_1 m_2}{m_1 - m_2}\right)(e^{m_1 t} - e^{m_2 t})$$

$$= CV\left(\frac{\alpha^2 + \omega^2}{\omega}\right)e^{-\alpha t}\sin \omega t.$$

Since $\int_0^\infty e^{-\alpha t} \sin \omega t \, dt = \frac{\omega}{\alpha^2 + \omega^2}$, $\int_0^\infty i \, dt = CV$

which is, of course, the final charge on the capacitor. For the total amount of energy dissipated in R_1 we could put

$$\int_0^\infty i^2 R_1 \, dt = c^2 V^2 R_1 \frac{(\alpha^2 + \omega^2)^2}{\omega^2} \int_0^\infty e^{-2\alpha t} \sin^2 \omega t \, dt$$

$$= C^2 V^2 R_1 \frac{\alpha^2 + \omega^2}{4\alpha} = C^2 V^2 R_1 \frac{1/L_1 C}{4R_1/2L_1} = CV^2/2$$

but this is unnecessary, since the quantity of electricity CV was all supplied at the voltage V, and the total energy A_1 supplied was therefore CV^2; of this $CV^2/2$ was stored in the capacitor, and the remaining $CV^2/2$ was dissipated in R_1. This shows, as we stated before, that the efficiency of charging the capacitor is 50%, whatever the value of R_1.

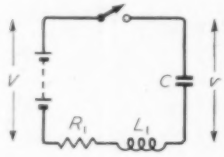

Fig. 1

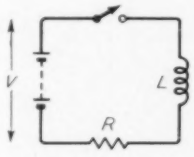

Fig. 2

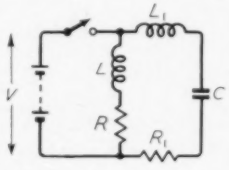

Fig. 3

In the simple case of Fig. 2 $i = \frac{V}{R}\left(1 - e^{-\frac{R}{L}t}\right)$.

The total amount of energy dissipated up to the time t_1 is

$$\int_0^{t_1} i^2 R \, dt = \frac{V^2}{R} \int_0^{t_1} \left(1 - 2e^{-\frac{R}{L}t} + e^{-2\frac{R}{L}t}\right) dt$$

$$= \frac{V^2}{R}\left(t_1 - 2\frac{L}{R} + \frac{L}{2R}\right) = \frac{V^2}{R}\left(t_1 - \frac{3}{2}\frac{L}{R}\right)$$

The first term is the energy that would have been dissipated if the current had jumped at once to its final value V/R. It is what van der Pol calls the pseudo dissipation W. In addition to the dissipated energy, the battery supplies the energy T stored in the magnetic field, viz., $\frac{1}{2}V^2 L/R^2$, so that the total energy supplied is

$$A_2 = V^2\left(\frac{L}{2R^2} + \frac{t_1}{R} - \frac{3L}{2R^2}\right)$$

of which the first term is the stored energy, the second term the pseudo dissipation which is proportional to t_1, and the last term the energy that was saved due to the initial current being smaller than its final value.

Combining Figs. 1 and 2 in the network of Fig. 3 we have for the total energy supplied

$$A = A_1 + A_2 = V^2\left(C - \frac{L}{R^2} + \frac{t_1}{R}\right)$$

Subtracting the pseudo dissipation $W = V^2 t_1/R$ we have $A - W = V^2(C - L/R^2)$

and since $U = \frac{1}{2}V^2 C$ and $T = \frac{1}{2}V^2 L/R^2$

$$A - W = 2(U - T)$$

in accordance with the network theorem.

To illustrate the theorem we have chosen a very simple example. In the publications of Lorentz and in the recent paper by Tellegen much more complicated problems are considered, such as the application of variable and alternating voltages to complex networks. Tellegen states one question as follows. Can anything be said about the distribution of energy that depends only on the admittance matrix of a $2n$-pole network and not on the particular network used for realizing it? He arrives at the following result. If on a linear, constant, passive, gyratorless $2n$-pole at rest, constant voltages are suddenly applied, then, when the final state has been reached, the total energy delivered to the $2n$-pole exceeds the energy representing the loss by dissipation at the final rate, supposed to start at once, by twice the excess of the electric over the magnetic energy. The difference between the magnetic and the electric energy will at any moment depend only on the admittance matrix of the $2n$-pole and not on the particular network used for realizing it. This being true at any moment, it is true for the mean values.

This interesting theorem, which we have illustrated by a simple example, is thus of very wide application.

G. W. O. H.

CATHODE-COUPLED VALVES

Graphical Method of Design

By T. W. Brady, B.Sc.

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Introduction

THIS paper will deal with a graphical method of solving cathode-coupled amplifier problems. By this term, Ross[1] and Lyddiard[2]—previous writers on the subject—meant the circuit shown in Fig. 1, in which one triode section acts as a cathode follower. The method to be described is quite general, however, and is demonstrated, applied to the circuit of Fig. 2.

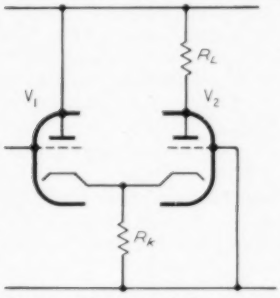

Fig. 1. *Simple cathode-coupled amplifier.*

First, it will be shown how the quiescent conditions of the amplifier in Fig. 2 may be represented on the anode characteristics of the valve. In this case the valve selected is a theoretically perfect Mullard ECC33, its characteristicsbeing identical with published data.

Secondly, it will be shown how the dynamic operating conditions may be deduced, together with some notes on output impedance, and thirdly the design of the amplifier will be considered.

Quiescent Conditions

By a trial and error solution of the amplifier in Fig. 2 using the published data it can be shown that the quiescent conditions are as follows:—

$$E_{HT} = 300\ \text{V}$$
$$I_{a1} = 2{\cdot}1\ \text{mA}$$
$$V_{g1} = -5\ \text{V}$$
$$I_{a2} = 5{\cdot}4\ \text{mA}$$
$$V_{g2} = -3\ \text{V}$$
$$I_{a1} + I_{a2} = 7{\cdot}5\ \text{mA}$$
$$V_k = 15\ \text{V}$$

A copy of the anode characteristics is now taken (Fig. 3). Since 15 V exists across the cathode resistor R_k, this leaves 285 V across each triode section with its associated load resistor. The load lines (for 20 kΩ and 40 kΩ) are drawn in from the 285-V point on the abscissa. These lines are marked "20 kΩ, V_2" and "40 kΩ, V_1" respectively for easy reference.

MS accepted by the Editor, July 1953

A cathode-bias line is now drawn for each triode section. This term designates a line giving the net bias or grid voltage on the section for various currents through the cathode resistor. For example V_1 has a 10-V positive bias applied, so that if 5 mA were to flow through the 2-kΩ cathode resistor the net bias or grid voltage would be zero. Note here that this current may not be due to V_1 alone, but to a combination of currents through V_1 and V_2. Nevertheless the statement still holds true. So on the $V_g = 0$ grid-voltage curve a mark is plotted against 5 mA. Similarly, if 6 mA were to flow through the cathode resistor, the net grid voltage would be -2 V. Therefore, on the $V_g = -2$-V grid-voltage curve a mark is plotted against 6 mA. Carrying the process further, a line is constructed and marked "V_1 Bias Line", for easy reference.

The whole process is repeated for V_2, giving a line higher than that for V_1 because of the higher applied positive bias. If by design the applied bias were the same for both triode sections, there would be only one bias line.

The next step is to draw composite load lines. These are lines which represent the sums of the anode currents of V_1 and V_2. Just as a normal load line gives the graphical determination of anode current when the grid voltage is known, the composite load line gives the sum of anode

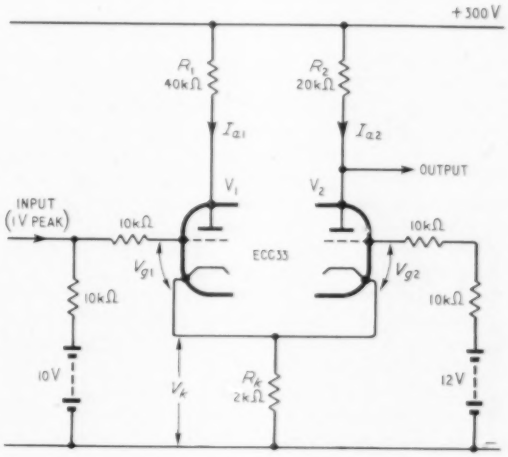

Fig. 2. *Form of amplifier treated in the text.*

currents for both sections when the grid voltages are known. However, since the grid voltages may be different for the two triode sections, these current sums when plotted must be referred to the grid voltage of one section and then to that of the other. The following description should make this clear.

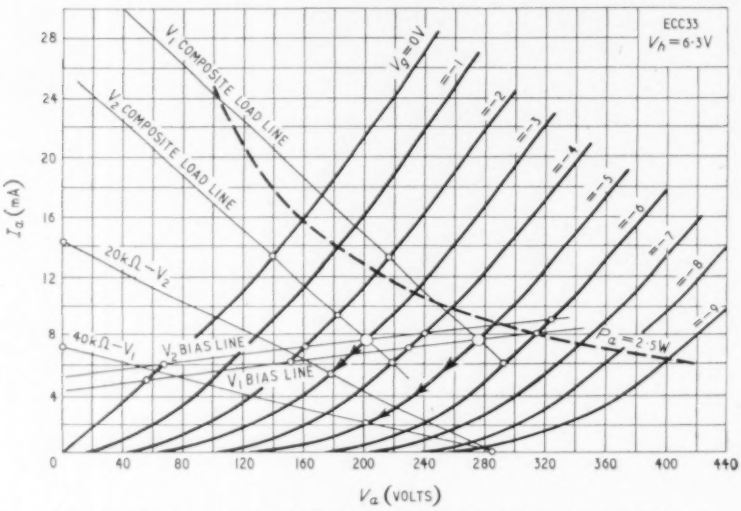

Fig. 3. *Construction for determining the quiescent operating conditions.*

Whatever the net bias or grid voltage of V_1 may be, the grid voltage of V_2 is 2 V greater because of the higher applied positive bias. Therefore the anode currents of V_1 at integral values of its grid volts are measured, and to these are added the anode currents of V_2 at corresponding values of grid volts. For example, the anode current of V_1 at $V_{g1} = -2$ V is 4 mA and the anode current of V_2 at $V_{g2} = 0$ is 9·2 mA. The sum is therefore 13·2 mA and this is plotted on the $V_g = -2$-V grid line, since $V_{g1} = -2$ V for this condition. Proceeding in this manner yields a composite load line for V_1.

The composite load line for V_2 follows by plotting the current sums obtained previously, at values of grid voltage 2 V greater than for V_1. This means the current sums are referred to V_2, by transposing them to the left by 2 grid volts on the graph.

Before proceeding further with the construction, a point must be explained in connection with the load lines and composite load lines. It has been assumed throughout that the voltage on the abscissa from which the load lines are drawn is 285 V. This is only true for the combination of anode currents which make a sum of 7·5 mA, so producing a drop of 15 V across the cathode resistor. Any increase in this current should cause the load lines to be drawn from a point on the abscissa to the left of the 285-V mark, while a decrease in current should cause the load lines to be drawn from a point to the right. The composite load lines would therefore move left and right of their present positions.

However, when the voltage drop across the cathode resistor is 15 V, the quiescent conditions must be along the load lines and composite load lines just drawn. In the construction so far, these lines must be regarded as devices to find information and not the loci of valve conditions. These will come later.

Coming back to the construction, the intersections of the composite load lines with the corresponding bias lines give the net grid voltage of each triode section, and also each intersection gives the total current through the cathode resistor, which must be the same in each case. (This is a check on the construction.) Considering each triode section in turn, the grid voltage curve must be followed down from each intersection of composite load and bias lines until the corresponding triode-section load line is encountered. This yields the individual quiescent conditions for the triode section. Thus for V_1 the conditions are 2·1 mA and 84 V across the 40-kΩ anode load resistor, and for V_2 5·4 mA and 108 V across the 20-kΩ load resistor. The net grid voltages are -5 V and -3 V respectively.

On reflection, it may now appear that a good deal of this construction is unnecessary. Since the voltage drop across the cathode resistor is known to be 15 V, the net grid voltages are immediately known to be -5 V for V_1 and -3 V for V_2. These points could have been plotted on the load lines at once and the rest of the construction ignored. However, composite load lines must be drawn to obtain dynamic operating conditions, and for the initial design of an amplifier.

Dynamic Conditions

Assume for the sake of argument that a positive signal of 1 V peak is injected into the grid of V_1. The instantaneous conditions are required at the limit of grid swing. These could be obtained by starting afresh and completely replotting, and

before discussing an easier way it may be as well to describe the essentially new steps.

V_1 has now effectively 11 V positive bias applied, and not 10 V as before. Therefore it needs a new bias line. Moreover, the composite load lines must be redrawn since the net grid voltage on V_1 must be 1 V less than V_2 and not 2 V as before. As an example the anode current of V_1 at $V_{g1} = -2$ V is 4·0 mA, and to this must be added the value of V_2 anode current at $V_{g2} = -1$ V, that is 7·8 mA. The total of 11·8 mA is plotted on the $V_g = -2$-V line for V_1 composite load line, and on the $V_g = -1$-V line for V_2 composite load line.

All of this work is not required, however (Fig. 4). It is known that V_2 bias line will not be changed in position since no signal has been injected on the grid of V_2. Therefore, the first step is to draw the new composite load lines and find where the new composite load line for V_2 meets the old (and still valid) bias line for V_2. This point is the new composite load–bias line intersection for V_2. Moving horizontally to the right until the new composite load line for V_1 is met gives the new intersection for V_1.

A further, or rather alternative, simplification is to draw the new composite load line for V_2 only, find the composite load–bias line intersection for V_2 as before, and then move horizontally to the right by a distance equivalent to 1 grid volt. Both of these methods are shown in Fig. 4.

Once the intersections are found, the grid voltage curves must be followed down to the load lines as before. The conditions obtained are:—

$$I_{a1} = 2{\cdot}6 \text{ mA}$$
$$V_{g1} = -4{\cdot}2 \text{ V}$$
$$I_{a2} = 5{\cdot}05 \text{ mA}$$
$$V_{g2} = -3{\cdot}2 \text{ V}$$
$$I_{a1} + I_{a2} = 7{\cdot}6 \text{ mA}$$
$$V_k = 15{\cdot}2 \text{ V}$$

Voltage across $R_1 = 104$ V.

Voltage across $R_2 = 101$ V.

The voltage gain from V_1 is therefore 20 and from V_2 about 7.

Two things will immediately be noticed about these results. First, I_{a1} and I_{a2} measured separately do not add up to $I_{a1} + I_{a2}$ measured at either (or both) of the composite load–bias line intersections. This is mensuration error. Second, the voltage across the cathode resistor is 15·2 V and not 15 V. Strictly speaking, therefore, the load lines are not based correctly, but since in practice it is very difficult to plot a voltage correct to 0·2 V on the abscissa of the anode characteristics, this fact may be ignored.

The dynamic load lines may not be the same as the steady-state or quiescent load lines as they were in the example. If such a case arises, the composite load–bias lines intersections and quiescent points on the load lines must be obtained first, the dynamic load lines are then drawn through the quiescent points (in the case of resistance loads) or based on the quiescent points for reactive loads. New composite load lines must be constructed for dynamic operation, but the procedure is as before.

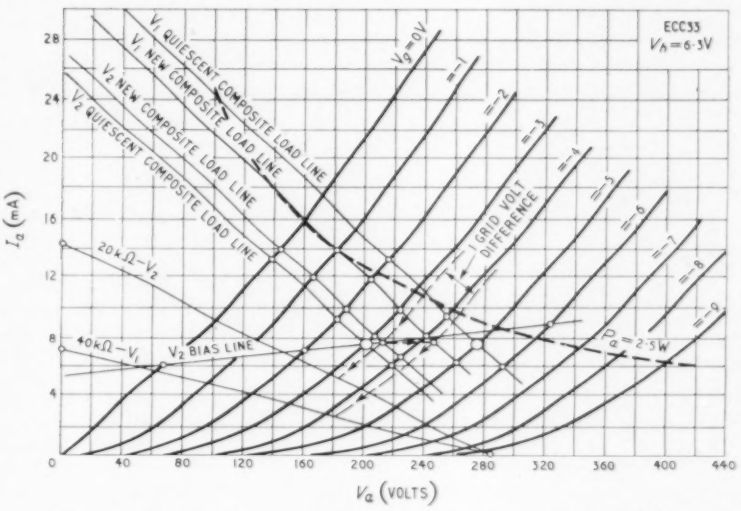

Fig. 4. *Construction for obtaining the dynamic operating conditions.*

An interesting construction follows from this. A vertical line drawn through a triode-section quiescent point represents zero dynamic load resistance. That would be physically a capacitor of infinite capacitance connected between the anode of the section (say V_2) and negative h.t. This, of course, is a dynamic short circuit at any frequency however low, and the construction may now be used to find the output impedance of the circuit if the output is taken from V_2.

To find the output impedance an arbitrary input signal is chosen, and the output voltage is

found by the construction demonstrated. The construction is then repeated with zero dynamic load resistance, and from this the increase in V_2 anode current is obtained. By dividing the output voltage by the increase in current the output impedance is calculated. Readers will recognize that an open-circuit and short-circuit test has been carried out (on paper), and this is the recognized way of measuring output impedance.

An analytical solution of the output impedance yields the formula:—

$$Z_{out} = R_2\left(\frac{[(\mu_2+1)(r_{a1}+R_1)+(\mu_1+1)r_{a2}]R_k + r_{a2}(R_1+r_{a1})}{[(\mu_2+1)(r_{a1}+R_1)+(\mu_1+1)(r_{a2}+R_2)]R_k + (R_1+r_{a1})(R_2+r_{a2})}\right)$$

To solve this formidable expression numerically, the parameters must be extracted from the valve anode characteristics, so the graphical method just enumerated is very much simpler and quicker.

Design

A problem arises in the application of this method to amplifier design, which is that since the value of voltage across the common cathode resistor is not known initially, the voltage from which to drawn the load lines is not known either. The practical solution is to draw the bias lines in first, since these do not alter with supply voltage, and then to draw the load lines from the value of supply voltage. In the previous construction, this of course was 300 V. One composite load line is now drawn, and this will give the net grid voltage on the triode section selected, usually within a fraction of a volt of its true value. Thus the drop in voltage across the common cathode resistor is known very nearly. As a practical example, a first cast at this problem gave a net grid voltage on V_2 of $-3{\cdot}3$ V; this meant that approximately 15 V existed across the cathode resistor, and this was then subtracted from the supply voltage to give 285 V, from which point the load lines were drawn. If a quick approximate solution is required, it is often possible to estimate the voltage across the common cathode resistor by drawing the load lines and bias lines only, since it is known that the quiescent point for a triode section must lie along its bias line, and also must lie to the right of the load line of smaller anode resistance. This follows, of course, because both composite load lines must lie to the right of the load line of smaller anode resistance.

It may be remembered that when dynamic conditions were obtained for the amplifier in Fig. 2, the change in voltage across the cathode resistor was neglected. This was justified at the time. In general, it is safe to neglect this change unless signals of the order of a few volts (or greater) are injected.

Conclusion

This method gives a solution to all cathode-coupled amplifier problems, without the necessity of drawing separate special curves or resorting to formulae. While it is not possible to refer immediately to a derived set of curves (vide Ross), amplifiers can be designed using this construction, as common RC-coupled amplifiers are designed by drawing simple load lines on the same valve anode characteristics.

Several variations of this scheme are possible, and it is quite easy to use a simpler construction if we are quite familiar with the theory. For example, it is not really necessary in Figs. 3 and 4 to draw the bias line and composite load line for V_1, since when the composite load–bias line intersection for V_2 is found, the corresponding point for V_1 can be found by moving horizontally to the right by the corresponding grid voltage difference.

The method has two disadvantages. First, it requires a trial solution to determine the voltage across the cathode resistor in the quiescent case, and secondly, it is difficult to determine small voltage changes accurately. However, since an actual valve performance is very rarely the same as the published data, small voltage changes deduced from a graph of any sort are only approximate to say the least.

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[1] S. G. F. Ross. "Design of Cathode-Coupled Amplifiers", *Wireless Engineer*, July 1950, Vol. 27, p. 212.
[2] J. A. Lyddiard. "The Cathode-Coupled Amplifier", *Wireless Engineer*, March 1952, Vol. 29, p. 63.

MEASUREMENT OF WAVEGUIDE IMPEDANCE

Use of Simple T Junction

A. Cunliffe, B.Sc., Ph.D., A.R.C.S., and **D. P. Saville**, M.Sc.

(University College of Hull)

SUMMARY.—The two symmetrical limbs of a right-angled T junction are connected to a waveguide section with a sliding-short termination and to the impedance under test respectively. The other limb is connected to the source of oscillations. Voltage is measured by means of a small pick-up probe at a fixed point in the limb connected to the impedance under test. The unknown impedance is found from observations of the voltage resonance which occurs when the sliding-short is adjusted. The apparatus can be used over a wide frequency band.

1. Introduction

WAVEGUIDE impedance is usually measured either by the standing-wave method or by the 'magic T' waveguide bridge. Both these methods, and particularly the standing-wave method, have been so successful that there has been a tendency for other possible methods to be overlooked completely or not considered in detail. The present method, which does not appear to have been previously considered, utilizes a resonance technique which is well known at lower frequencies. Like the standing-wave indicator, the apparatus can be used over a wide frequency band.

The resonance technique involved is that in which a circuit is adjusted to resonance by altering a known standard component. For example, in the Hartshorn and Ward method for the measurement of dielectric constant[1] and in the Q meter, the system is tuned to resonance by adjusting a standard capacitor. In the present adaptation, the standard component takes the form of a section of waveguide with a movable short-circuiting termination. The input reactance or susceptance of this waveguide section can be calculated theoretically from its length and the characteristic impedance of the waveguide. Theoretically, if losses are negligible, it can be made to vary continuously from minus infinity to plus infinity by sliding the short-circuiting termination along the length of the waveguide.

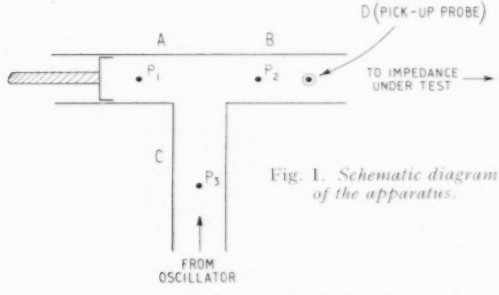

Fig. 1. *Schematic diagram of the apparatus.*

MS accepted by the Editor, June 1953

The arrangement is shown in Fig. 1. The sliding short-circuit moves in one of the symmetrical limbs, A, of a simple right-angled T junction, and the other symmetrical limb, B, is terminated with the impedance under test. Power enters the junction via the third limb, C, and a voltage pick-up probe is placed at a fixed point, D, in the limb B.

2. Theory of the Method

The method depends for its action on the use of *characteristic points*, the idea of which was introduced by Allanson, Cooper and Cowling.[2] Let P_1, P_2 and P_3 be characteristic points in the limbs A, B and C respectively. It is assumed that the waveguides transmit freely only the fundamental mode, and that P_1, P_2 and P_3, as well as the sliding-short, pick-up probe, impedance under test and the oscillator, are sufficiently far away from the junction for the evanescent higher-order modes formed there to have decayed to a negligible amount. It is also assumed that the junction itself contained between the points P_1, P_2 and P_3 is free from loss.

Let the source, as seen from P_3 and looking away from the junction, be equivalent to an ideal voltage source, E, with an internal series impedance, Z_3. Let the impedances in the other two limbs, referred to the points P_1 and P_2, and looking away from the junction, be Z_1 and Z_2 respectively. Then it can be shown that the voltage at either of the points P_1 or P_2 can be written as V where

$$V = \frac{\alpha}{\beta}\frac{E}{Z_3}\Big/\left(\frac{\alpha^2}{\beta^2 Z_3} - \frac{1}{\beta^2} + \frac{1}{Z_1} + \frac{1}{Z_2}\right) \quad . . \qquad (1)$$

and where for a given frequency of oscillation, α and β are constants, the squares of which are pure reactances. (1) can be written as

$$V = i/(A_s + A_1 + A_2) \quad . . \quad . . \quad . . \qquad (2)$$

where $i = \frac{\alpha}{\beta}\frac{E}{Z_3}$, $A_s = \frac{\alpha^2}{\beta^2 Z_3} - \frac{1}{\beta^2}$, $A_1 = \frac{1}{Z_1}$ and $A_2 = \frac{1}{Z_2}$.

Now a system comprising a constant-current source, i, with an internal shunt admittance, A_s, further shunted by admittances A_1 and A_2, as shown in Fig. 2, has a voltage, V, across it which can be calculated from an equation identical with (2). The circuit of Fig. 2 is, therefore, an exact equivalent circuit of the waveguide system. A_1 and A_2 are admittances referred to P_1 and P_2 of Fig. 1 respectively, looking away from the junction, and V is the voltage at either P_1 or P_2. If V can be measured, by making A_1 a known variable susceptance, the circuit can be used to measure A_2. Put $A_1 = jY_1$ and $A_2 = G_2 + jY_2$.

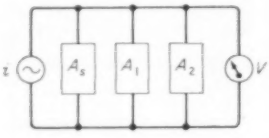

Fig. 2. *Equivalent circuit of Fig. 1—parallel resonance.*

If L_1 is the distance of the sliding-short from P_1, Y_1 can be calculated in terms of L_1. Neglecting losses,

$$Y_1 = -\frac{1}{Z_0}\cot\frac{2\pi}{\lambda_g}L_1$$

where Z_0, the characteristic impedance, and λ_g, the wavelength in the waveguide, can be calculated from its dimensions.

The unknown susceptance can be found as follows. Let Y_1' be the value of Y_1 for which V is a maximum with A_2 present. Let Y_1'' be the value of Y_1 for which V is a maximum with A_2 removed. Then

$$Y_2 = Y_1'' - Y_1' \quad .. \quad .. \quad .. \quad (3)$$

With the waveguide apparatus, to make the measurement with A_2 removed, A_2 is to be replaced by a quarter-wavelength of guide with a short-circuit termination.

To determine the unknown conductance, G_2, three methods are available:

(a) Measure $\Delta Y_1'$ and $\Delta Y_1''$, the changes in Y_1 needed to tune the system between the half-power points with A_2 present and with A_2 removed respectively. Then

$$G_2 = \tfrac{1}{2}(\Delta Y_1' - \Delta Y_1'') \quad .. \quad .. \quad .. \quad (4)$$

(b) Let V' and V'' be the resonant values of V with A_2 present and with A_2 removed respectively. Then

$$\frac{G_s}{G_s + G_2} = \frac{V'}{V''} \quad .. \quad .. \quad .. \quad .. \quad (5)$$

where G_s is the conductive part of A_s. If G_s is known, G_2 can be found by measuring V' and V''.

(c) Having found Y_2, use the apparatus to measure the susceptance Y_2' of A_2 as seen through a quarter-wavelength of guide. Then

$$G_2 = \sqrt{-\frac{Y_2}{Z_0^2 Y_2'} - Y_2^2} \quad .. \quad .. \quad (6)$$

This method has the advantage that the voltmeter is only required to indicate resonance, and therefore need not be calibrated.

In the above discussion, V is the voltage at the characteristic points P_1 or P_2. Now as far as variation in the position of the sliding-short only is concerned, the voltage at any point not too close to the junction along the limb containing P_2 will be proportional to V. This means that the same values will be obtained for Y_1', Y_1'', $\Delta Y_1'$ and $\Delta Y_1''$ with a pick-up probe placed at any point not too close to the junction along the limb containing P_2. Therefore, as far as these measurements are concerned, the pick-up probe need not be at the characteristic point P_3. This is important, because the positions of the characteristic points vary with frequency, and had it proved necessary to have the pick-up probe exactly at the characteristic point, the apparatus, once the hole had been drilled for the probe, would only be usable at a fixed frequency.

In a similar way, the probe need not be at the characteristic point if method (c) is used to measure G_2. It must, however, be at the characteristic point for method (b).

The hole for the probe and the depth of insertion of the probe should be small, to avoid distortion of the field. Rather more distortion is allowable when the probe is exactly at the characteristic point, because the distortion effectively forms an admittance in parallel with A_s. If the probe is not adjusted during a measurement, this admittance may then be considered to be part of A_s.

The positions of the characteristic points can be calculated theoretically, but they can more easily be found experimentally. Thus the sliding-short will be an integral number of half-wavelengths from P_1 when it is adjusted until there is no voltage at the pick-up probe. The characteristic points along the two symmetrical limbs will, of course, be symmetrically arranged with respect to the centre of the junction.

If the values of E, Z_1, Z_2 and Z_3 are E', Z_1' Z_2' and Z_3' when referred to points P_1', P_2' and P_3' distant $\lambda_g/4$ away from the points P_1, P_2 and P_3 respectively, it can be shown that the current at the point P_2' can be written as i' where

$$i' = -\frac{\alpha}{\beta}E' \Big/ \left(\frac{\alpha^2}{\beta^2}Z_3' - \frac{Z_0^2}{\beta^2} + Z_1' + Z_2'\right) \quad (7)$$

Equation (7) indicates a *series* resonance method for *impedance* measurement. The equivalent circuit is shown in Fig. 3. Z_1' is a known variable reactance. Since, as far as variation of Z_1' only is concerned, the current i' at P_2' is proportional to the voltage at any point not too close to the junction along the limb containing P_2, it follows that the apparatus can be used for the determination of impedance by a series-resonance method. That is, the same apparatus can be used either for admittance measurements with the circuit of Fig. 2, or for impedance measurements with the circuit of Fig. 3. In the admittance measurements, the admittance under test is to be referred to the characteristic point P_2. In the impedance measurements, the impedance under test is to be referred to the point P_2' distant $\lambda_g/4$ away from P_2.

3. Experimental Work

To test experimentally the efficacy of the method, a T junction was made from rectangular $\frac{1}{2}$-in. by 1-in. brass waveguide. The junction was an H-plane junction; that is, the plane of the T was parallel to the magnetic vector and the long side of the waveguide. The limb of the junction, which was at right angles to the other two limbs, was coupled, via an attenuator to provide "padding", to a CV 129 oscillator. One of the other limbs contained the sliding short-circuit, which was a plunger making mechanical contact with the waveguide wall a distance $\lambda_g/4$ in front of the short. The plunger was driven by means of a micrometer head. The remaining limb was terminated with a coupling flange for connection

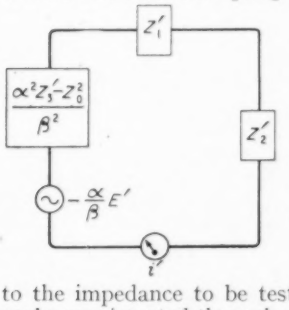

Fig. 3. *Equivalent circuit of Fig. 1—series resonance.*

to the impedance to be tested, and the pick-up probe was inserted through a hole drilled through the centre of the long side of the waveguide at a point along this limb. This hole had a diameter of 0·220 cm and was 6·294 cm from the centre of the junction. This was not a characteristic point for the free-space wavelength of 3·200 cm at which the measurements described below were carried out. The pick-up probe had a diameter of 0·125 cm and was tunable, the energy picked up being fed to a detector unit with a square-law characteristic.

To test the apparatus as an admittance-measuring device, it was used to measure the susceptance of a length of waveguide with a sliding-short termination similar to the one incorporated in the apparatus. The susceptance measured was then calculated theoretically, and experimental and theoretical results were compared. The positions of the characteristic points in the limbs containing the sliding-shorts were found by the method described in the last section. The short on the waveguide section being measured was then set at an arbitrary point and the other short adjusted until the detecting apparatus indicated resonance. The distances L_1 and L_2 of the shorts from their respective characteristic points were then obtained from the micrometer head settings. At resonance

$$Y_s + Y_1 + Y_2 = 0$$

where Y_s is the susceptance part of A_s.

$$\text{i.e.,}\quad Y_s - \frac{1}{Z_0}\left(\cot\frac{2\pi}{\lambda_g}L_1 + \cot\frac{2\pi}{\lambda_g}L_2\right) = 0 \qquad (8)$$

For most of the experiments, the pick-up probe was not actually inserted into the waveguide, but was withdrawn so that the tip of the probe was approximately level with the waveguide wall.

Fig. 4 shows a graph of $\cot 2\pi\, L_1/\lambda_g$ plotted against $\cot 2\pi\, L_2/\lambda_g$. That is, the normalized susceptance Z_0Y_1 is plotted against the normalized susceptance Z_0Y_2. The graph shows a straight line of slope -1 as required according to equation (8). The intercepts give a normalized susceptance Z_0Y_s equal to $-0{\cdot}24$. The table below gives some results for numerically-large normalized susceptance values.

TABLE

$Z_0Y_1\cot\frac{2\pi}{\lambda_g}L_1$	4·74	3·75	14·6	23·9	−19·8	−5.22
$Z_0Y_2\cot\frac{2\pi}{\lambda_g}L_2$	−5·24	−3·92	−16·8	−27·3	18·2	4·74

There is increasing uncertainty regarding the susceptance values as the numerical value of the normalized susceptance increases. This limitation arises because the short circuits are then very close to the characteristic points. The percentage errors in the measurement of L_1 and L_2 are then great, and also for these very high susceptance values, the effects of the finite conductivity of the waveguide wall become important. Thus a waveguide section an integral number of half wavelengths long can never form an exact short circuit. Losses due to the finite conductivity of the waveguide wall have been ignored in the theory, but they are not peculiar to this particular method.

As an example of the measurement of a

conductance, results are given below for the measurement of the input admittance of a section of waveguide containing a variable vane-type attenuator adjusted to a nominal value of

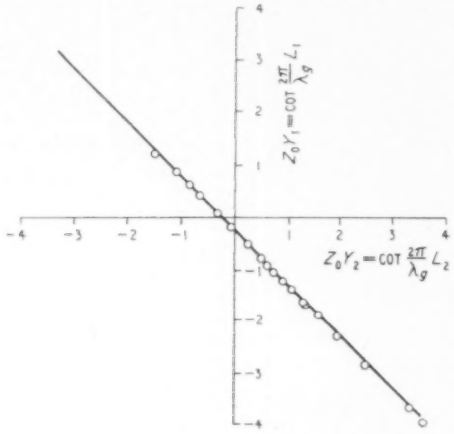

Fig. 4. *Graph of normalized susceptance in arm A plotted against normalized susceptance in arm B.*

25 db. The conductance measurement was made by the half-power method. The values obtained were as follows:—

$$Z_0G_s = 0{\cdot}66 \qquad Z_0Y_s = -0{\cdot}24$$

$$Z_0(G_s + G_2) = 1{\cdot}82 \qquad Z_0(Y_s + Y_2) = -0{\cdot}36$$

From these equations, $Z_0G_2 = 1{\cdot}16$ and $Z_0Y_2 = -0{\cdot}12$. Hence $Z_0A_2 = Z_0(G_2 + jY_2) = 1{\cdot}16 - 0{\cdot}12j$, from which it can be seen that the attenuator, as adjusted, formed a load which was approximately matched to the waveguide.

The results obtained seem to indicate that the accuracy of the method compares quite favourably with that of other methods at these wavelengths. The apparatus is easy to make, and the simplicity of the operating technique possibly offers scope for its use in industry.

Acknowledgments

Thanks are due to Professor L. S. Palmer for his interest and for providing the research facilities. D. P. Saville is indebted to the Council of the University College of Hull for the award of a Research Studentship.

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HARMONIC DISTORTION AND NEGATIVE FEEDBACK

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(University of Southampton)

THE effect of negative feedback on distortion has lately been the subject of a good deal of attention.* It has been known for many years that the use of negative feedback gives an improvement in distortion which is not accurately expressed by the factor $(1 + A\beta)$. This fact is to be expected since the gain A is not clearly defined in non-linear conditions. In addition, negative feedback reduces harmonics generated by the amplifier but produces new harmonics, so that even the term 'distortion' cannot be clearly defined.

In his article R. O. Rowlands gives a quantitative expression for the improvement of distortion by feedback. His most important conclusion is given in the statement that distortion due to a flat part of the V_o/V_i curve is not improved by negative feedback. While this is true in the case discussed by Rowlands, it is not generally true.

In the following, a detailed treatment is given of many possible cases and the effect of negative feedback is analysed for each.

At the start it is important to define the gain of an amplifier. When conditions are linear there is no ambiguity in the normal definition. In plotting the graph of output voltage against input voltage we can take the ratio of instantaneous output voltage to instantaneous input voltage, or we can take, for an assumed sinusoidal input, the ratio of the amplitudes of signal frequency. Both definitions give the same value of gain.

When the amplifier is non-linear the conditions are no longer so simple. The ratio of instantaneous output voltage V_o to instantaneous input voltage V_i depends now on the magnitude of the input voltage and can, therefore, be defined only for a given input voltage. But, for a sinusoidal input, the ratio V_o/V_i for maximum V_i is usually not the same as the ratio of the amplitudes of the fundamental frequency.

*R. O. Rowlands, "Harmonic Distortion and Negative Feedback", *Wireless Engineer*, June 1953; correspondence *Wireless Engineer*, September, October, November 1953.

MS accepted by the Editor, January 1954

Referring to Fig. 1, curve 1, assume that $V_o = aV_i - cV_i^3$ and $V_i = E \sin \omega t$, then for $V_i = E$, we have $V_o/V_i = a - cE^2$ but, since $\sin^3 \alpha = \frac{3}{4} \sin \alpha - \frac{1}{4} \sin 3\alpha$, it follows that the ratio of the amplitudes of angular frequency ω is $a - \frac{3}{4} cE^2$; this shows that one has to be careful when defining gain in a non-linear amplifier.

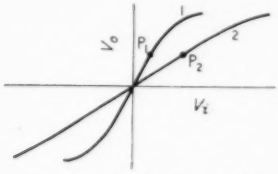

Fig. 1. *Curves representing amplifier characteristics* (1) *without feedback and* (2) *with feedback.*

The two definitions of gain give, however, the same result, if we apply them to the incremental gain dV_o/dV_i, at any point of the V_o/V_i curve. Assume, for instance, that in Fig. 1, the curves 1 and 2 give the relation between V_o and V_i without and with feedback respectively. Then for points on the curves with identical output (P_1 and P_2) the amplifier itself works under the same conditions; hence the incremental gain

$$\frac{dV_o}{dV_i} \text{ (at } P_2) = \frac{dV_o}{dV_i} \text{ (at } P_1)/(1 + A\beta),$$

A being the incremental gain at P_1. If, in curve 1, we express V_i as a function of V_o

$$\left(\text{i.e., if } V_i = F(V_o) \text{ and if } \frac{dV_i}{dV_o} = \frac{1}{A} \right)$$

we may write

$$V_i = \int_0^{V_o} \frac{1}{A} \, dV_o$$

where A may now be considered to be a function of V_o. Then, in curve, 2,

$$V_i = \int_0^{V_o} \frac{1 + A\beta}{A} \, dV_o = F(V_o) + \beta V_o.$$

The result, naturally, follows from the definition of feedback and shows clearly the relation between the two curves. If the input required for a given V_o is V_i in curve 1 and V_i' in curve 2, the above formula gives

$$\frac{V_i'}{V_i} = \frac{V_i + \beta V_o}{V_i} = 1 + \beta \frac{V_o}{V_i}.$$

Now V_o/V_i is the average slope A_{av} of curve 1. Thus it is seen that the average slope A_{av} of a V_o/V_i curve is reduced by feedback in the ratio $1 + A_{av}\beta$. (The fact is stated by F. G. Kerr in his letter in *Wireless Engineer*, September 1953.) It is, therefore, reasonable to use the average slope to define the gain of a non-linear amplifier, although this slope does not give the ratio of the amplitudes of fundamental frequency. In practice this fact should not matter, because only in cases of excessive distortion will the latter ratio differ appreciably from A_{av}.

It is easy graphically to construct the V_o/V_i curve with feedback when the curve without feedback is given. Of several methods the following is particularly suitable.

Let us assume that we draw from the origin a straight line of some convenient slope A; for a given output V_o the required input may be written $V_i = F(V_o) = \frac{V_o}{A} + \delta V_i$. Then δV_i is the horizontal distance of the V_o/V_i curve from the straight line $V_o = AV_i$; i.e., δV_i is the amount by which the actual input required for an output V_o differs from that of the ideal amplifier which obeys the relation $V_o = AV_i$. With feedback we have

$$V_i' = F(V_o) + \beta V_o = V_o \frac{1 + A\beta}{A} + \delta V_i;$$

this shows that the horizontal distance of the new V_o/V_i curve with feedback is the same as before when measured from the straight line of slope $\frac{A}{1 + A\beta}$.* Hence if we reduce the V_i-scale by the factor $1 + A\beta$, the line given by $V_o = \frac{AV_i}{1 + A\beta}$ stays as before, while the new V_o/V_i curve may be obtained directly by reducing the horizontal distances from the straight line by the factor $1 + A\beta$. This is done in Fig. 2. The slope of the straight line is chosen to be equal to the average slope of the V_o/V_i curve (this is particularly convenient) and the value $1 + A_{av}\beta$ is assumed to be equal to 2.

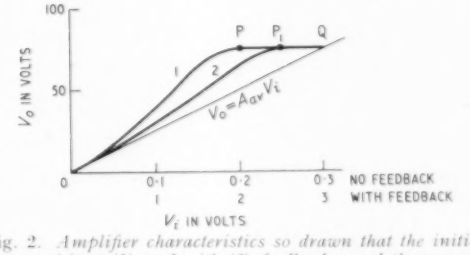

Fig. 2. *Amplifier characteristics so drawn that the initial slopes without* (1) *and with* (2) *feedback equal the average slope.*

For the curve 1 without feedback the initial slope, at $V_i = 0$, and the average slope are identical; hence they are also identical for the curve 2 with feedback. Point P on the old curve becomes P_1 on the new; Q stays at its place. It is easily seen that the reduction of distortion is marked even for the flat part of the curve.

The point is brought out more forcefully in the example shown in Fig. 3(a). This type of V_o/V_i

*Compare p. 132 of Mr. Rowlands's article.

curve may be obtained by two valves working in class C push-pull. The amplifier is to be so driven that the effective input voltage, with or without feedback, does not exceed ± 1V. Then the distortion is caused only by the flat part between $-0{\cdot}5$ V and $+0{\cdot}5$ V.

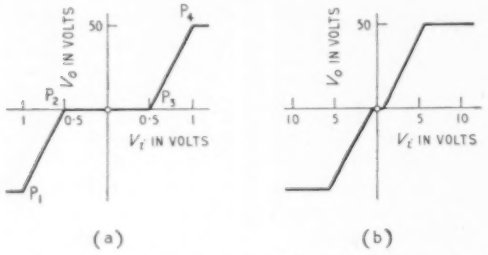

Fig. 3. *Characteristic of push-pull class C amplifier without feedback (a) and with feedback (b);* $1 + A\beta = 10$.

Let us assume that negative feedback is applied of magnitude $1 + A\beta = 10$ where A refers to the incremental gain on the steep part of the curve. In order to reach the cut-off point P_3 the required input voltage is $0{\cdot}5$ V as before, since feedback is not yet effective. To reach point P_4 an additional input of 5 V is required. The curve with feedback hence looks as shown in Fig. 3(b), the V_i-scale being contracted by the factor 10. It is seen that feedback cannot alter the flatness of the curve between P_2 and P_3, but for small angles of no-current flow it reduces by a factor approaching $1 + A\beta$ the fraction of the cycle during which the flat part is operative. Thus feedback tends to change class C into class B amplification. Fourier analysis shows that in the graph of Fig. 3(b) the magnitude of the 3rd harmonic is approximately proportional to the fraction of the cycle during which no-current flows, so that the improvement is of the order of $1 + A\beta$.

If, as shown in Fig. 4(a), the slope between $-0{\cdot}5$ V and $+0{\cdot}5$ V is not zero, but still much smaller than that of the other parts, negative feedback acts in two ways. First, it reduces the ratio of the two slopes, secondly it reduces the fraction of time during which the smaller slope is operative; both effects tend to reduce distortion. The V_o/V_i curve with feedback is shown in Fig. 4(b); $1 + A\beta$ is 10 on the steep part of Fig. 4(a). With a sinusoidal input, the shape of the output curve, with and without feedback, may be seen from Fig. 5. In (a) $\theta = \sin^{-1} 0{\cdot}5 = \frac{\pi}{6}$; in (b) $\theta = \sin^{-1} \frac{1}{6} = 0{\cdot}17$. The magnitude of the output of fundamental frequency in Fig. 5(b) is approximately 20% larger than in (a); this may be verified by Fourier analysis, which also shows that the ratio of 3rd harmonic to fundamental in Fig. 5(a) is 0·27 and 0·036 in Fig. 5(b). The improvement is not much less than $1 + A\beta$, A referring to the steep part. As will be seen from the examples, the factor $1 + A_{av}\beta$, gives a reasonable approximation throughout.

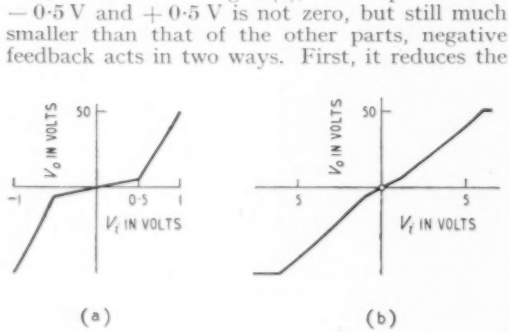

Fig. 4. *Amplifier characteristic having two parts of different slopes (a) and with feedback (b);* $1 + A\beta = 10$.

In all the previous cases it has been assumed that the V_o/V_i curve is symmetrical about the origin and that the effective input, and hence the maximum output, are the same with or without feedback.

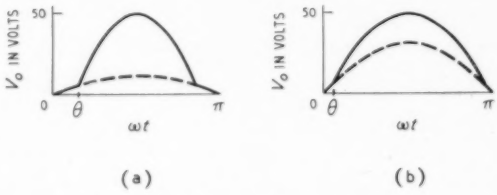

Fig. 5. *Effect of amplifier of Fig. 4. on a sine wave; (a) without and (b) with feedback.*

In Fig. 6 the case is different. Negative feedback will reduce the slope for positive input voltage while the slope for negative voltage remains zero. With a sinusoidal input the output is a sine curve with the negative portion cut-off; hence negative feedback brings no improvement.

In Fig. 7 the V_o/V_i curve is also asymmetrical about the origin, but now negative feedback gives a definite improvement. The gain for positive input voltage is assumed to be A_1, that for negative voltage A_2, where $A_2 < A_1$. The output consists of a positive and a negative half-sine curve; the ratio of the amplitudes is A_1/A_2. The distortion may be measured by the ratio of the difference in amplitudes to the larger amplitude. Hence without feedback the distortion is $\frac{A_1 - A_2}{A_1} = \frac{\delta A}{A_1}$. With feedback the distortion is

$$\left[\frac{A_1}{1 + A_1\beta} - \frac{A_2}{1 + A_2\beta}\right] \Big/ \left(\frac{A_1}{1 + A_1\beta}\right) = \frac{\delta A}{A_1} \frac{1}{1 + A_2\beta}$$

The positive output is the same with and without feedback, the negative is given by curves 1 and 2 respectively (Fig. 8).

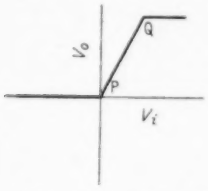

Fig. 6. *Discontinuous characteristic of a single-sided amplifier.*

The last two cases, which may be compared with that dealt with by Rowlands, differ essentially from those described in Figs. 1–4. The cause of this difference must be seen in the fact that in Figs. 6 and 7 the drive is not the same with and without feedback. Let us, for example, suppose that in Fig. 7 the effective positive drive is 1 V with and without feedback, and that with feedback an input of 5 V is required for the purpose. This means that $A\beta = 4$ for positive input while $A\beta = 2$ for negative input, the slope being half that in the positive region. Hence with feedback the effective negative drive becomes $\frac{5}{3}$V, which is $\frac{2}{3}$V more than without feedback. The output of fundamental frequency, which is equal to the arithmetic means of the positive and negative amplitudes, is increased by 22%. This should be taken into account when assessing the improvement caused by feedback.

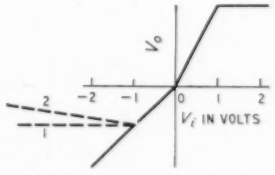

Fig. 7. *Single-sided amplifier characteristic having two portions of different slopes.*

If the V_o/V_i curve of Fig. 7 were flat for a negative drive exceeding 1 V (dotted line 1) the result would be quite different. With feedback the positive output is given by curve 1 as before; but the negative flat part is reached when the instantaneous input voltage is -3 V. Hence the amplifier is driven another 2 V into the negative flat part, and the total negative drive becomes 3 V. The shape of the negative part of the output curve is given initially by 2 and then by 3 of Fig. 8. Even without applying Fourier analysis one can see that negative feedback cannot be expected to bring about much improvement and may even make things worse.

A large increase in distortion would be caused by negative feedback if the slope of the V_o/V_i curve were negative for a negative drive exceeding 1V (dotted line 2). The negative part of the output curve would then be given in Fig. 8 by a combination of lines 2 and 4, the effective negative drive being about 4·2 V.

From the various cases discussed the following conclusions can be drawn:—

1. V_o/V_i Curve Symmetrical about the Origin

The effective drive can be considered to be the same with and without feedback. The improvement in distortion is then, with reasonable accuracy, given by the factor $1 + A_{av}\beta$.

2. V_o/V_i Curve Asymmetrical about the Origin

(*a*) *The average gain the same for the positive and negative parts* (*Fig.* 9). From what has been said on page 119 it is clear that the effective drive and hence the maximum positive and negative output will be the same with and without feedback. Feedback will, therefore, again give an improvement similar to that with symmetrical V_o/V_i curves.

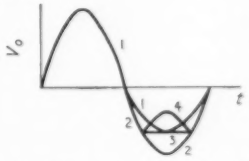

Fig. 8. *Waveforms obtained with the different forms of amplifier characteristic of Fig.* 7.

(*b*) *The average gain different for the positive and negative parts.* The effective drive with feedback will usually be the same for the part having the larger average gain, but it will be larger for the part with the smaller gain. The effect of feedback on distortion depends then on that part of the V_o/V_i curve which, lying on the side with the lower average gain, extends beyond the point to which the amplifier is driven in the absence of feedback. This means that, if the V_o/V_i characteristic is expressed as a power series, the effect of feedback will sometimes be determined by the terms of high powers.

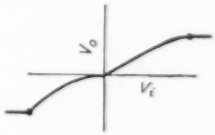

Fig. 9. *Characteristic having the same average gain for its positive and negative parts.*

The way in which distortion should be assessed seems to be largely a matter of taste. For curves of similar type the deviation from a straight line will give a fair indication. Often the V_o/V_i curve may not contain a clearly defined straight part, and in this case it is probably best to draw a straight line through the origin and the maximum V_o on the side having the larger average gain, and to use this line for reference. If this is done in Figs. 5 and 6 the improvement thus found approximates to that calculated for the 3rd harmonic.

AUTOMATIC TUNING FOR PRIMARY RADAR

By S. Ratcliffe, B.Sc.

Introduction

THIS article summarizes the results of a period of research into the basic techniques of automatic frequency correction (a.f.c.) for primary radar, particularly the problems arising in airborne microwave systems. Basic principles of good a.f.c. design are emphasized, and no attempt is made to give a catalogue of circuits. Some of these principles, such as the need to use 'designable' circuits, have already been stated very explicitly elsewhere,[1] but the importance of this point still does not seem to be sufficiently appreciated.

An obvious problem in the design of a microwave radar is that set by the very high ratio of the transmitted frequency to the receiver bandwidth. For example, 4 Mc/s is a typical bandwidth for a 10,000-Mc/s radar, the equivalent of a broadcast receiver bandwidth of, say, 400 c/s in 1 Mc/s. For this reason, the need for some form of a.f.c. was recognized very early in the history of microwave radar. This was a major factor influencing the choice of the reflex klystron[2] for development as the receiver local oscillator.

Initially, it was usual to provide a receiver bandwidth considerably greater than the optimum, in order to allow for possible imperfections in the a.f.c. system, but with increasing demands on radar performance arising from higher speeds of aircraft, the subsequent loss in overall efficiency becomes intolerable. Efforts have accordingly been made to improve the design to the point where it is possible to use a receiver bandwidth not far from the theoretical optimum.

It is essential to notice a fundamental difference between the applications of a.f.c. to broadcast and radar equipments. In the broadcast problem there are many receivers to one transmitter, and it is economical to employ an elaborate arrangement to maintain the transmitted frequency constant. In a primary radar there is one transmitter to each receiver, and it is normally sufficient to control the frequency of either the transmitter or the receiver local oscillator so as to maintain the desired intermediate frequency, the absolute frequency being of less importance. In practice the receiver frequency is controlled, because of the relative ease with which this can be done. Because of this choice between complicating the transmitter or the receiver, their designs and that of the a.f.c. system are inter-related.

The frequency variations with which the radar a.f.c. may have to deal can be divided into two categories. First, there are the long-term variations, such as those due to changes in ambient temperature and pressure. Secondly, there are the variations in magnetron frequency due to changes in the matching during rotation of the scanner, and similar effects. It will be shown that the phase of the reflection from the radome, for example, varies so rapidly that the magnetron frequency may change by a megacycle or so between one pulse and the next. The speed with which the frequency varies from these causes is considerably greater than that of the long-term fluctuations previously mentioned, and an a.f.c. technique which is otherwise adequate may require drastic modification if these short-term effects are important.

There is a further difference between broadcast a.f.c. and the corresponding control in a radar receiver. The former has error information continuously available from the discriminator, so that the a.f.c. system can be designed by the technique developed for negative-feedback amplifiers. In a radar a.f.c. system, however, information about the tuning error is only available while the transmitter is firing (i.e., for a small fraction of the operating time) so that conventional negative-feedback theory must not be too hastily applied to such a system.

Earlier radar a.f.c. systems have a reputation for unreliability. One of the reasons for this is that fault-finding is difficult; since, when the a.f.c. is faulty, the signal on which it should operate is often also missing in consequence. The difficulties are increased by the fact that the setting-up may involve as many as five preset adjustments which interact with each other and are mostly frequency dependent. It was necessary to provide many of these adjustments because much of the design was arrived at empirically, without time to establish the effect on performance of production tolerances on components. The result is that equipments were often so designed that an acceptable performance could only be obtained by selecting a local-oscillator valve to suit a particular set. The maintenance and production problems can be greatly simplified by the use of 'designable' a.f.c. circuits, in which the effect of production

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tolerances can be predicted and minimized and in which the need for preset adjustments can be greatly reduced, as can the necessity for much of the routine 'testing' of this apparatus. Such tests as are necessary can usually be made without the need for special test gear, such as pulsed-signal generators, which are needed for the correct maintenance of many earlier a.f.c. systems.

One major contribution to the problem of eliminating preset adjustments is the use of a broad-band a.f.c. system in which frequency sensitive adjustments are avoided. This involves some departure from the earlier a.f.c. techniques, in which the local-oscillator frequency was controlled only by variation of the voltage applied to one of its electrodes. The range over which the frequency can be controlled by such means is limited, and it is also necessary to provide a preset tuning adjustment with which long-term corrections are made. Apart from the disadvantage of adding to the number of adjustments, this method lowers the receiver efficiency. Unless the electronic-tuning range is more than adequate to cope with all the long-term frequency drifts, a.f.c. will often fail, but increasing the electronic-tuning range brings a number of difficulties. This technique tends to be particularly unsatisfactory at higher frequencies where the electronic-tuning range may well be inadequate even to cope with frequency drifts in the local oscillator itself. To avoid limitations of this kind, it is necessary to introduce a mechanical servo which can control the frequency by variation of the physical dimensions of the resonator of the oscillator. A variety of suitable techniques is now available.

In the design of a magnetron it is necessary to strike a compromise between a high power output and a low 'pulling figure'. As this pulling figure increases, the receiver bandwidth must be increased to allow for the effects of scanner pulling. This increase in bandwidth results in a deterioration in the receiver sensitivity and the design must aim at a suitable compromise. The use of an a.f.c. system having a sufficiently rapid response to deal with scanner pulling will make possible a magnetron design having a higher pulling figure, with a consequent increase in efficiency, or a receiver having a narrower bandwidth and increased sensitivity. In the next section is discussed the relation between these design parameters, and the conditions under which it may be economical to use a fast a.f.c. system.

A.f.c. should be regarded and designed as an integral part of the radar system, and not simply as an addition to an existing collection of components which is used to deal with the problems left unsolved by the designers of the scanner, magnetron, local oscillator, etc.

Scanner Pulling and Design of Radar Systems

Reference has already been made to the fact that the frequency of a magnetron depends, to some extent, on the impedance into which it is working. Before discussing the behaviour of the magnetron in more detail, it will be of interest to consider some of the relevant properties of the scanner.

In the sense used here, this consists of a waveguide run, with one or more bends, one or more rotating joints, the aerial proper, and, usually, a radome which protects the aerial from the elements. Mechanical tolerances on the components set a limit to the accuracy of impedance matching that can be achieved, and on the X-band, for example, a voltage standing-wave ratio (v.s.w.r.) of 0·85 to 0·9 is probably the best that can be guaranteed in production. For the present purposes we are not so much concerned with this mismatch as with the way in which it varies as the aerial rotates, since it is this variation which influences the a.f.c. problem.

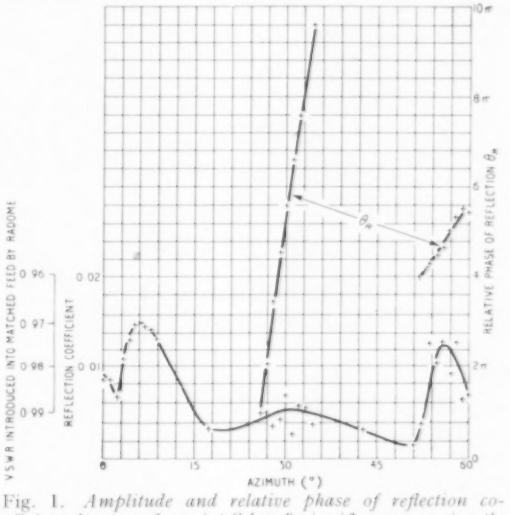


Fig. 1. *Amplitude and relative phase of reflection coefficient due to radome (at S-band);* $|e_R|\theta_R$ *versus azimuth. Radome = 1/12-in. glass-fibre laminate.*

Fig. 1 shows the variation in the reflection from a typical aircraft radome as a function of the direction of the beam. It will be noted that the phase of the reflection varies at a remarkably high speed with change of azimuth. This is because the radome is large compared with the wavelength, and hence a small percentage change in the path length can be responsible for a large shift in the phase of the reflection.

The variation in the reflection from a rotating joint will arise from eccentricity or asymmetry of the rotating section, and although no precise data is available, it does not appear likely that any

rapid change in phase will occur during rotation. For this reason, it is probably justifiable to neglect the variation in the frequency pulling due to a well-designed rotating joint.

The output power of a magnetron is extracted by means of a coupling to the resonant cavity. Consequently, the frequency at which a magnetron oscillates depends on the impedance presented to it by the external load. The relationship between this impedance and the magnetron frequency and power output is commonly shown on a Rieke diagram,[3] such as Fig. 2, which shows lines of constant frequency and of constant power output on a circle diagram. Additional lines may be sketched, if necessary, by interpolation. The peak-to-peak variation in frequency as the operating point is taken round the circle for the reflection coefficient $K = 0{\cdot}2$ (v.s.w.r. $= 0{\cdot}67$) is termed the 'pulling figure' of the magnetron. To a first approximation, the effect of a small mismatch can be predicted by assuming the pulling to be directly proportional to the reflection coefficient. Pulling figures of the order of $0{\cdot}2\%$ of the transmitter frequency are usual. To keep down the pulling figure it is usual to undercouple the load, sacrificing some of the output power in the process. This has happened in Fig. 2, for example.

The Rieke diagram gives the magnetron fre-

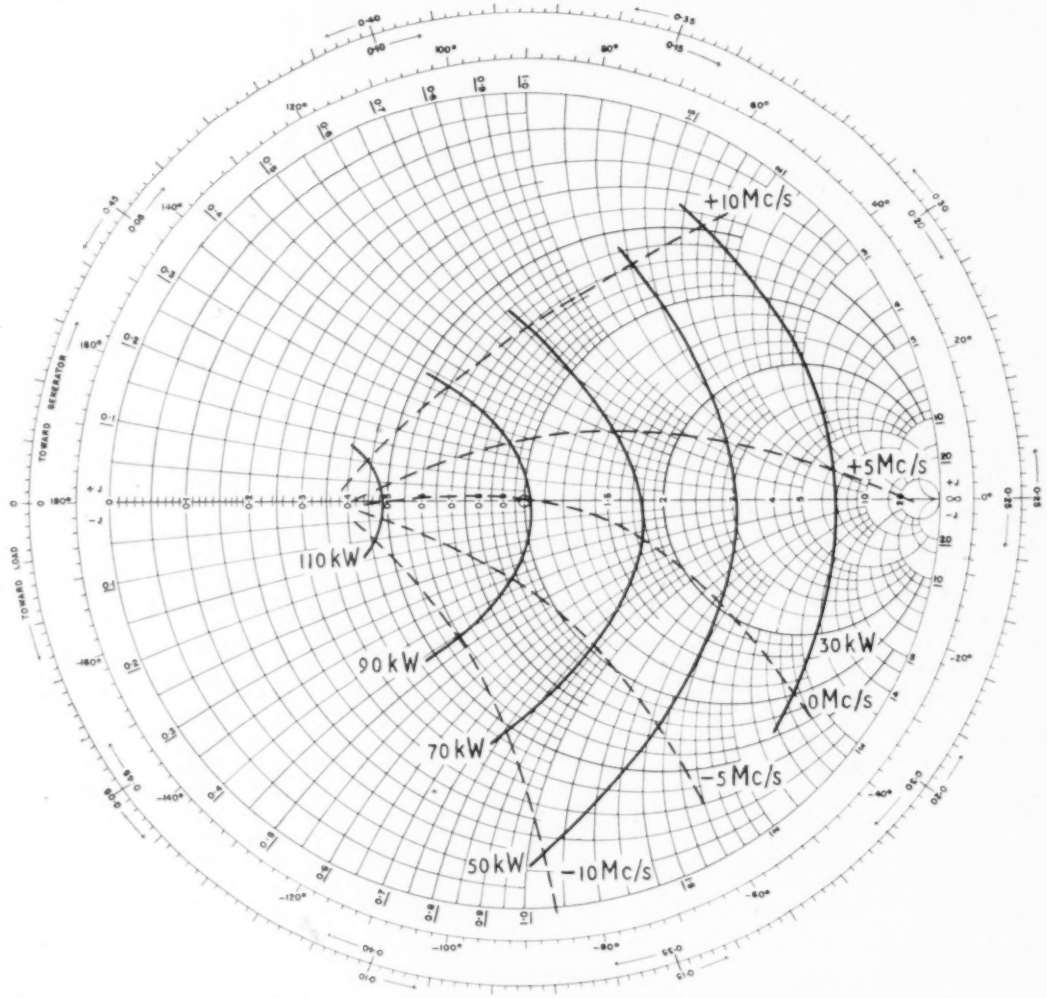


Fig. 2. *Rieke diagram of 4J50 magnetron (6300 gauss, 10 A), showing lines of constant frequency and power.*

quency as a function of the impedance presented to the magnetron by the load. The aerial is usually separated from the magnetron by a guide many wavelengths long, and the intervening length of waveguide will modify the characteristics of the system. This usually has only a second-order effect on the scanner pulling and need not concern us here.

The diagram of Fig. 2 gives the steady-state frequency only. At the beginning of a pulse there is no energy from the scanner returning to the magnetron which, therefore, behaves as though it were feeding a matched load. The determination of the transient variations in frequency of the magnetron is difficult but, for airborne installations at least, the waveguide is usually sufficiently short to make it possible to assume that the magnetron spectrum is centred about the steady-state frequency.

The efficiency of a magnetron is generally limited by the need to restrict the pulling figure. Before discussing the factors influencing the choice of this figure, we require to know the relationship between pulling figure and power output. The pulling figure is controlled by the ratio of the transformer between the output line and the magnetron cavity. Within limits the effect of any such change can be deduced from the Rieke diagram, such as Fig. 2, by multiplying all the impedances by the appropriate factor and re-drawing the diagram. From the modified Rieke diagrams we can determine the way in which the efficiency of a magnetron of this type would vary if the pulling figure were changed. The results are summarized in Fig. 3.

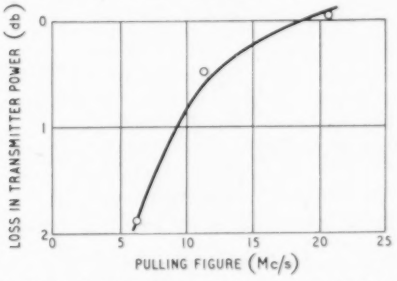


Fig. 3. *Output power of magnetron versus pulling figure.*

Failing a perfect scanner, a certain amount of frequency pulling must occur during the scanner rotation, and the design of the whole equipment should take account of this fact. If the receiver bandwidth is increased in order to reduce the sensitivity of the receiver to changes in the frequency of the received signals, this will affect the overall performance for, as the bandwidth of a radar receiver is increased, the fraction of the transmitted energy spectrum which the receiver can accept will increase steadily, but eventually the noise in the receiver will be increasing at a greater rate. Thus the signal-noise ratio will suffer. The effect of varying receiver bandwidth on the visibility of small signals on a c.r.t. display has been investigated by Haeff,[4] for an A-scope and by Payne-Scott[5] for a P.P.I. Broadly speaking, these writers agree that the optimum i.f. bandwidth is given by the reciprocal of the pulse width, and that for a bandwidth y times this, the sensitivity is reduced by a factor $4y\ (1 + y)^{-2}$. This is plotted in Fig. 4. The second-order effect of i.f. bandwidth on the receiver noise factor is neglected.

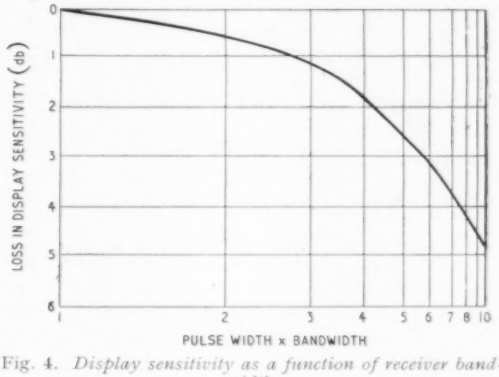


Fig. 4. *Display sensitivity as a function of receiver bandwidth.*

We also require to know the effect on the response of a receiver of given bandwidth of any change in the received frequency. It is assumed, for the present purposes, that we are dealing with an ideal rectangular pulse, that the receiver response curve is also rectangular, and that the sensitivity of the equipment is directly proportional to the received energy. Then, if the receiver is correctly tuned, the output power will be given by

$$cP_0 = \int_{-x}^{+x} \frac{\sin^2 \pi u}{\pi^2 u^2}\, du = 2 \int_0^x \frac{\sin^2 \pi u}{\pi^2 u^2}\, du = 2I_x, \text{ say,}$$

where c is a constant, and x is equal to half the product of receiver bandwidth and the pulse width. If m denotes the product of a tuning error with the pulse width, then the output power, for this tuning error, will be given by

$$cP = I(x + m) + I(x - m)$$

and the loss in signal power will be given by

$$\frac{P}{P_0} = 1 - \frac{I(x + m) + I(x - m)}{2I_x}$$

A graph of I_x is given in Fig. 5 and a graph of P/P_0 as a function of m for $x = 1, 2, 3$, and 4, is given in Fig. 6, and from either of these the effect of a given amount of mistuning can be determined.

Hence it is possible to choose the correct receiver bandwidth to handle a pulse having a known amount of frequency modulation due to pulling.

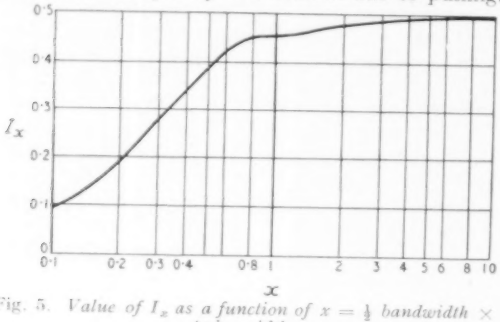


Fig. 5. *Value of I_x as a function of $x = \frac{1}{2}$ bandwidth × pulse width.*

In the presence of any appreciable amount of frequency modulation, some amplitude modulation of the receiver output must ensue, particularly as a practical i.f. amplifier will not have a perfectly flat-topped response, as assumed in Fig. 6. The designer must decide what amount of such amplitude modulation he is prepared to tolerate, bearing in mind that this modulation can only be reduced by lowering the performance of the whole equipment in other respects. The choice may be influenced by factors such as the following:—

(a) It may be known that the mismatch due to the radome will be worst in some special direction. This often happens in an airborne installation when the shape of the radome is determined by aerodynamic considerations. The requirements for the system may be such that it is not essential to maintain the peak performance in this direction.

(b) A signal-comparison technique is often used to give a sensitive indication of the bearing of a target. Fig. 1 shows that a small change in the direction of the beam from an aerial may produce a considerable change in the phase of the reflection, and hence produce a significant change in the magnetron frequency. The resulting amplitude modulation of the receiver output will introduce an error in the apparent bearing of the target, and it will therefore be necessary to choose a receiver bandwidth which will keep this error down to a tolerable amount.

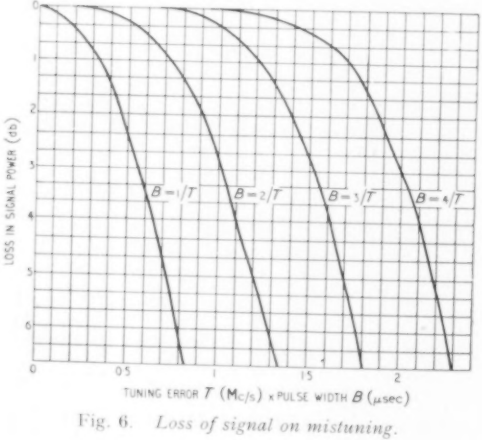


Fig. 6. *Loss of signal on mistuning.*

From an estimate of the probable variation of the scanner impedance during rotation, it is possible to predict the resultant scanner pulling for a given magnetron-pulling figure. It can be assumed, for small reflections, that this is given by

Scanner pulling = (Variable component of reflection coefficient) × (Magnetron-pulling figure) × 5

(This assumes that the pulling varies linearly with the reflection coefficient and that the pulling figure is measured for a reflection coefficient of 0·2).

From the various relations given above, it is possible to deduce the receiver bandwidth required for any given magnetron-pulling figure, and to determine the optimum combination of bandwidth and pulling figure.

Suppose it is desired to design a system using a magnetron of the 4J50 type, and a radome having properties similar to those plotted in Fig. 1. Suppose the pulse width is to be 1·0 μsec, and that one can tolerate a loss in performance of 2 db due to mistuning of the receiver at the positions of the scanner producing the worst mismatch, which will be taken as corresponding to a reflection coefficient of 0·015 (v.s.w.r. = 0·97). The pulling due to this effect will be ($F \times 0{\cdot}075$ Mc/s) where F is the magnetron-pulling figure.

Table 1 shows, for various pulling figures,

(1) the receiver bandwidth required,
(2) the loss in display sensitivity which results,
(3) the loss in transmitter power due to under-coupling,
(4) the sum of items (2) and (3).

TABLE 1

Pulling Figure (Mc/s)	(1) Bandwidth (Mc/s)	(2) Loss in Receiver (db)	(3) Loss in Magnetron (db)	(4) Overall loss in Efficiency (db)
5	1	0	2·5 approx.	2·5
10	1·5	0·2	1	1·2
15	2·4	0·8	0·5	1·3
20	3·2	1·3	0	1·3

There is little point in considering the use of a pulling figure appreciably greater than 20 Mc/s, since the operating point is rapidly moving into the region where magnetron operation is uncertain and, unless the radar has a very short waveguide run, trouble will be experienced from the 'long-line' effect.[6]

For this example, at least, if the receiver bandwidth is correctly chosen to suit the magnetron-pulling figure, the value of the latter is not critical. The value given in column (4) for the overall loss in efficiency, 1·3 db for the existing magnetron, represents the loss in performance due to the imperfection of the radome. This loss would be considerably reduced if the pulse were narrower and vice versa, since it is the ratio of the scanner pulling to the width of the magnetron spectrum that determines the loss in receiver performance due to the fall in the limiting sensitivity of the display.

If the pulse-width is reduced, the mean power output of the transmitter must, of course, be kept constant to avoid loss in performance due to the necessary increase in receiver bandwidth.

The above discussion has omitted any reference to the a.f.c. system. If this were capable of following instantaneously the variations in magnetron frequency due to the radome, the loss in performance due to the latter could be avoided. The difficulty is that even for the 10-cm aerial shown in Fig. 1, the rate-of-change of phase of the reflection is about π radians per degree azimuth. Now a change of π radians in the phase of the reflection may change the magnetron frequency by the whole amount of the pulling, and if the beamwidth is, say, 6° and the radar p.r.f. has been chosen to put six pulses on the target at each sweep, we have one pulse per degree of azimuth, and the a.f.c. must follow a transmitter which may jump in frequency by the whole amount of the pulling between one pulse and the next. We will discuss later the possibility of designing an a.f.c. system which, when faced with this situation, can make possible a significant reduction in the receiver bandwidth.

Source of Control Information

A schematic diagram of a cm-wave radar a.f.c. system is shown in Fig. 7. The requirement is to maintain a constant separation between the frequency of the transmitter and that of the local oscillator, this difference being the intermediate frequency of the receiver. To this end, samples of the output of the transmitter and of the oscillator are applied to a mixer. The resulting difference signal is amplified and applied to a discriminator. The output of the discriminator is proportional in magnitude and sense to any departure of the difference frequency from the desired value and can thus be used to actuate a control system which reduces the tuning error.

For various reasons[7] it is now general practice to derive the a.f.c. discriminator input from a mixer and i.f. amplifier distinct from those used to handle the incoming radar signals.

The design of this mixer need not be discussed at great length, since most of the problems are common to the design of signal mixers. The main difference in emphasis arises because of the importance of maintaining close tolerances on the gain round the a.f.c. loop, and therefore on the output of the a.f.c. mixer, even at the expense of the mixer sensitivity.

Care must also be taken with the method of extracting a sample of the transmitter power, to ensure that a small percentage harmonic content of the transmitter output does not become the dominant signal at the a.f.c. mixer.[7]

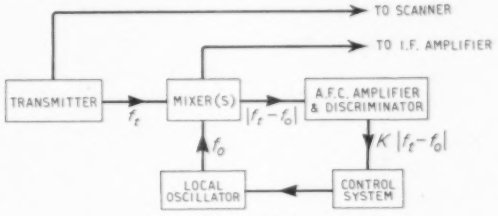


Fig. 7. *A.F.C. system.*

In many a.f.c. systems difficulty is experienced from the harmonics of the i.f. which are generated by the crystal mixer. Thus, if the local-oscillator frequency differs from that of the transmitter by half the desired i.f., the second harmonic produced by the crystal may be sufficient to actuate the a.f.c. circuits, which will then hold the receiver tuned to the wrong frequency. To minimize this trouble, it is necessary to choose with some care the operating conditions of the a.f.c. mixer. The even harmonics generated in the mixer (and spurious signals due to transmitter pulse currents) can be greatly reduced by the use of a balanced a.f.c. mixer.

The level of local oscillator and transmitter power at the mixer will probably be of the order of 2 mW and $\frac{1}{2}$ mW for satisfactory harmonic performance, giving an i.f. output of about 0·1 V. There are a number of arguments influencing the decision as to which of the two signals should be made the larger.

(a) It may be difficult to obtain 2 mW from the local oscillator, or to keep down leakage from the transmitter to a level low compared with $\frac{1}{2}$ mW.

(b) The r.f. impedance of the crystal will be mainly determined by the larger of the two signals. If this signal is the c.w. from the local oscillator, the impedance presented to

the waveguide will be substantially the same during the transmitter pulses as during the intervals between them. If the transmitter drive is the larger, the crystal impedance will change during the pulse, and this may complicate the problem of designing the 'plumbing' so as to keep down interaction between the a.f.c. and signal mixers. This change of impedance also complicates the problem of ensuring that the a.f.c. gets the desired level of l.o. power during the transmitter pulse.

(c) Troubles due to the video components of the mixer current will be reduced by keeping the pulsed-signal low.

(d) It is important to maintain close tolerances on the gain round the a.f.c. loop. The output from the a.f.c. mixer will depend mainly on the smaller of the two input signals, and it is desirable, therefore, to choose as the greater of the two signals that likely to vary most widely. The argument is further complicated, however, by the fact that, as shown later, the reflector characteristics of a klystron are such as to cause an increase in the loop gain as the oscillator is mistuned electrically away from the frequency at which it gives maximum power output. This effect is reduced if the local-oscillator signal is made the smaller of the two, since the falling-off of the i.f. output at the edges of the a.f.c. range will to some extent compensate for the increasing slope of the reflector characteristic.

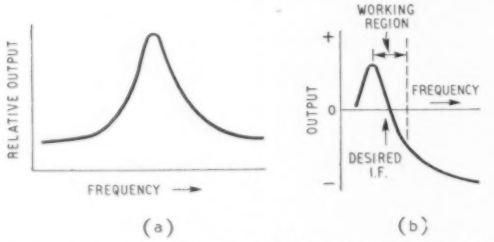


Fig. 8. *Response of single-tuned circuit (a) and output of detector (b).*

In practice, it is usual to operate the mixer with a large transmitter sample and low local-oscillator drive. The mixer output is fed through one or two amplifier stages to a discriminator which then develops an output proportional to the difference between the i.f. and its desired value. The minimum usable i.f. and discriminator bandwidth is determined by the need to accommodate within the pass-band a substantial portion of the spectrum of the transmitted pulse. If this is not done, the a.f.c. may lock to one of the side-lobes of the spectrum. C. Baron, of R.R.E., Malvern, has shown that a discriminator peak separation of at least 2/(pulse width) is desirable to avoid these troubles. A wider separation is sometimes used in order to keep up the 'pull-in' range. The design of the i.f. amplifier stages is conventional and need not be discussed here. Typical designs will be seen in various published accounts of a.f.c. systems.[7]

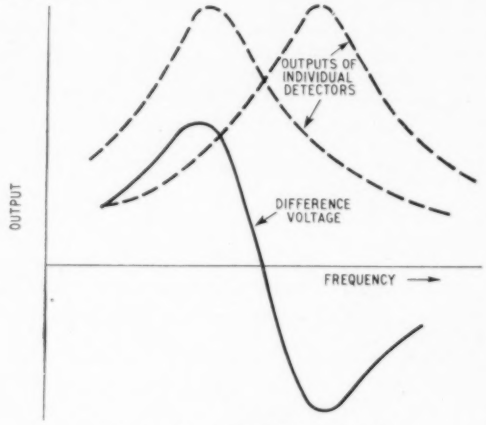


Fig. 9. *Response of discriminator.*

The discriminators are identical in principle with those used in broadcast receivers and a brief summary only will be given here of the various possible circuits. Discriminator circuits can be divided into two classes:—

(a) Those using the frequency response curve of a single-tuned circuit. If a circuit is tuned to a frequency somewhat different from the desired i.f. then the output from a detector connected to it will vary with frequency in the manner shown in Fig. 8(a). If the output corresponding to the desired i.f. is backed-off by a suitable opposing voltage, the curve of Fig. 8(b) is obtained. This output is capable of providing a.f.c. information but it has the disadvantage that the position of the cross-over will vary with the strength of the input signal.

The Round-Travis[8] circuit is the parent of many commonly used discriminator designs. It consists essentially of two circuits each coupled to a detector. One circuit is tuned to a frequency above and the other below the desired i.f. and the frequencies are so chosen that the outputs are equal at the desired i.f. If the two outputs are then subtracted from each other, the difference voltage will vary with frequency in the manner shown in Fig. 9. Among the many possible ways of applying

this idea is the 'Weiss' discriminator,[7] often wrongly described as a variant of the Foster-Seeley discriminator.

(b) The Foster-Seeley or 'phase' discriminator uses a band-pass coupling with a centre-tapped secondary. The phase of the voltage induced in the secondary varies with frequency being, at resonance, 90° different from that of the primary voltage. The phase of the secondary voltages is compared with that of the primary with the aid of two diodes in a bridge circuit as shown in Fig. 10. The Foster-Seeley circuit ceases to be very satisfactory when the peak-separation required is much more than 10–15% of the i.f. because a tight coupling between primary and secondary is then required, and it becomes difficult to control this coupling with sufficient accuracy, or to preserve the symmetry. For this reason, it is now usual to employ the Weiss discriminator[7] for wider peak-separations.

The i.f. signals, from which the discriminator output is derived, consist of a succession of pulses having a duty cycle of the order of 1 in 1,000, and an error voltage of this form is not very suitable for controlling the oscillator frequency. A common solution is to make the discharge time-constant $R_L C$ of the diode load in Fig. 10 long compared with the pulse width, so that the capacitor, which is charged during the pulse through the diode, is discharged much more slowly by R_L. The voltage across the diode load thus decays slowly between pulses, giving an output voltage that is more suitable for use in the control circuit. This method of lengthening the pulses introduces several complications, among which are:—

(a) The detection efficiency of the diodes is very low because of the need to charge up a large load capacitance during the pulse. This means that more i.f. gain is needed if the signal level at the diode outputs is not to be too low. It is commonly necessary to use two parallel-input i.f. stages feeding the pair of circuits forming a Round-Travis discriminator in order to supply an adequate charging current for these capacitors.

(b) Because of the use of load capacitors which cannot be charged by the diodes in the period of the pulse, the diodes present very heavy damping to the tuned circuit of the discriminator. This damping is virtually unpredictable and will depend on the relative values of the tuning error during the preceding pulses, because of charges remaining on the capacitors from one pulse to the next.

Tests on the response curve of the discriminator conducted with a c.w. source will offer little guidance as to the performance of the a.f.c. system in its normal role. Nor do tests conducted with a pulsed-signal generator correctly simulate working conditions, since given a steady error the diodes may succeed in charging up their load capacitors to some reasonable fraction of the incoming r.f. voltage, but this situation will not arise when the a.f.c. is operating normally. The effective damping may therefore be appreciably different.

(c) The output from such a discriminator will be a function of the integral of the tuning error, as well as of the error itself. In addition, if the error changes rapidly, only one of the diodes in Fig. 10 will be able to conduct. The other will be biased-off by the charge remaining on the load capacitor from the previous error voltage. Thus the discriminator output voltage will underestimate a large and rapidly changing error. These two effects greatly complicate the task of predicting the performance of an a.f.c. system in which this type of discriminator is used.

It is much better, therefore, to separate the two functions of rectifying and stretching the pulses. This makes possible a simpler design, gives a predictable performance, makes it possible to use a normal c.w. signal generator for the setting up, and reduces the amount of expensive i.f. amplification required.

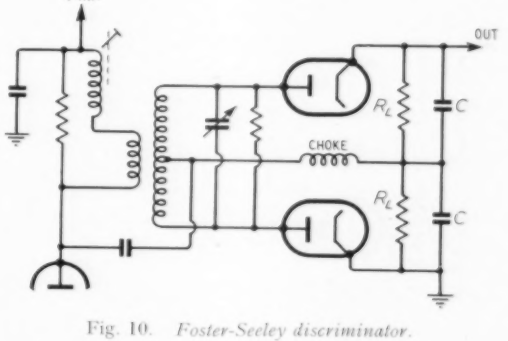


Fig. 10. *Foster-Seeley discriminator.*

A fundamental problem that seems to have received little attention concerns the behaviour of any discriminator, and therefore of any a.f.c. system, when the spectrum of the transmitted pulse differs appreciably from the theoretical. For example, the current delivered to the magnetron from the matching transformer will normally 'droop' to some extent during the pulse, and this 'droop' will result in a certain amount of frequency modulation, usually termed 'pushing'. It now becomes difficult to assign any clear

meaning to the term 'magnetron frequency', or to state precisely the conditions under which the receiver is said to be 'on tune'.

Ideally, an operator performing the tuning operation by hand would set the tuner to the point at which small signals were most clearly seen above the background noise. In practice, there may not be a suitable signal available, and there will be human errors, but the a.f.c. is handicapped by the fact that the discriminator cannot use this criterion and must attempt to judge the correct tuner setting by some indirect means.

A normal type of a.f.c. with a linear discriminator will seek a tuner position such that the average frequency of the mixer output coincides with the desired i.f., the average being taken over a band of frequencies roughly equal to the peak separation of the discriminator, but with some weight given to the part of the spectrum lying within the skirts of the discriminator characteristic. Qualitative argument shows that, if this band over which the average is taken is roughly equal to the bandwidth of the radar receiver proper, this criterion is a sensible one, and should result in a good approximation to the optimum performance. The difficulty is that, for reasons previously mentioned, it is otherwise desirable to use a peak separation appreciably greater than the optimum receiver bandwidth. Thus, given a badly asymmetrical spectrum, some compromise may be necessary. On the other hand, the development problem of improving the transmitter must be assessed in relation to any proposed complication of the a.f.c. to deal with the poor spectrum.

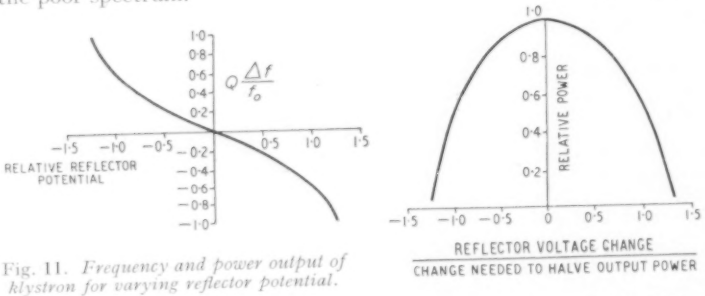


Fig. 11. *Frequency and power output of klystron for varying reflector potential.*

Means of Controlling the Oscillator Frequency

Having obtained, from the discriminator, information about the tuning error, this must be used in some way to correct the local-oscillator frequency. There are two important ways in which the frequency of a reflex klystron, or similar oscillator, can be varied.

(a) By mechanical variation of the dimensions of the resonant circuit.

(b) Electronic tuning, in which the reactance presented to the resonant cavity by the electron stream is varied by changes in the electrical conditions of the klystron. Deliberate frequency changes are normally produced by variation of the reflector voltage. A negligible time delay is incurred in producing a frequency change in this way, and most war-time radars used an a.f.c. system which controlled the oscillator frequency by means of the reflector voltage.

It is not appropriate in this article to give a detailed discussion of the theory of the reflex klystron such as is available elsewhere.[2,9] It will be sufficient to quote some results obtained from first-order theory.[9] Fig. 11 shows the output and frequency of a klystron as a function of reflector potential. The electronic tuning range (Δf) between half power points is given by

$$\Delta f = \frac{1{\cdot}2\, f_0}{Q} \qquad (2)$$

where f_0 is the radio frequency of the oscillator, and Q is the loaded-Q of the resonant cavity. The slope of the reflector-voltage versus frequency characteristic is also inversely proportonal to Q, and increases as the operating point moves away from the centre of the mode. Fig. 12 shows the electronic-tuning rate as a function of the deviation of the reflector voltage from the optimum value.

In general, klystrons designed for use with reflector a.f.c. use a low-voltage high-current (i.e., low impedance) beam, which couples to a low-Q cavity, giving a wide electronic-tuning range. Where the wide electronic-tuning range is not required, it is common to use a high-voltage low-current (high impedance) beam coupled to a high-Q cavity.

The frequency of the klystron may be varied by mechanical means either by changing the volume of the cavity ('inductance' tuning), or by changing the grid spacing ('capacitance' tuning) or both. The choice of tuning method has considerable influence on the performance of the klystron. Capacitance tuning produces a large frequency shift for a very small mechanical motion, and is often employed in valves in which the resonant cavity is inside the valve envelope, in order to simplify the problem of transmitting the mechanical motion. There is, however, a limit to the range over which capacitive tuning can be used without seriously impairing the efficiency of the oscillator, since when the frequency is raised by increasing the gap length, the transit angle through the gap

increases rapidly, because the electrons have further to go and less time in which to make the journey. A further difficulty is that a portion of the resonator wall must be flexible. This leads to microphony troubles due to sound waves striking the diaphragm or, if the diaphragm is stiff, to mechanical hysteresis effects or, not infrequently, to both.

If inductance tuning is employed, a relatively large mechanical motion is required. The available tuning range is much greater, since the gap length is constant.

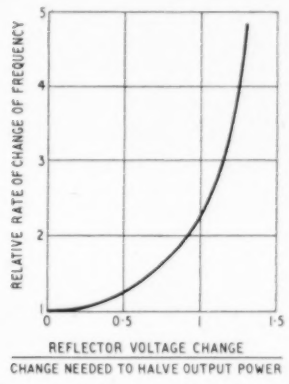


Fig. 12. *Slope of electronic tuning characteristic.*

In view of the limited range over which an a.f.c. system can tune a klystron by varying its reflector voltage, it is becoming more usual to attempt some form of control which can vary the dimensions of the resonant cavity used with the valve and which is therefore able to apply much larger corrections. The tuning mechanism must be light and simple, and capable of varying the oscillator frequency without introducing serious backlash problems. One attempted solution, to which considerable ingenuity has been devoted, is the thermally-tuned klystron. Valves of this type, which are now of little more than historical interest, were developed both in the U.S.A. and in this country.

Writers at the end of World War II painted a rosy picture of the thermally-tuned valve and its advantages. In practice, these are outweighed by two major difficulties: the manufacturing problems, and the microphony problem. The conventional low-voltage klystron is already sufficiently difficult to develop and produce, and the insertion of what is virtually another valve into the same envelope complicates still further the production problem, and greatly increases the amount of research necessary to develop valves for new frequency bands. The microphony troubles arise because of the extreme inefficiency of the thermal tuner strut regarded as an electric motor. Because of the small mechanical force available to perform the tuning operation, the mechanism must be very light and flexible. The microphony troubles then become more or less inevitable. Interested readers should consult the various published accounts[2,7] or, for the later work in this country, the patent literature.[10]

In view of these difficulties, other forms of mechanical frequency control, in which a more conventional and efficient motor mechanism is fitted outside the valve envelope, are now generally being adopted.

The outstanding mechanical problem involved in such a technique is that of designing a tuning mechanism having sufficiently low backlash. The rate of change of frequency with the position of a tuning plunger (if inductive tuning is used) or of a flexible diaphragm (if capacitive tuning is used) will almost certainly exceed 100 Mc/s per cm and may be as high, on X-band, as 20,000 Mc/s per cm. If the tolerable backlash cannot exceed 1 Mc/s or so, the design of a suitable mechanism presents some difficulty. A number of techniques are available, however, which make it possible to supplement the frequency control exercised by this motor by a fine control using the reflector characteristic. Under these circumstances, the mechanical requirements are not so stringent, and a satisfactory design can fairly readily be achieved. To avoid the mechanical problems entailed in breaking the mechanical coupling between the l.o. and the tuning motor when valve replacement is necessary, it is now usual to use one of the 'plug-in' klystrons recently developed for this work. With these valves, much of the resonant cavity and the entire tuning mechanism form a permanent part of the equipment. The valve itself consists of an envelope bearing a pair of contacts which engage with the rest of the resonator.

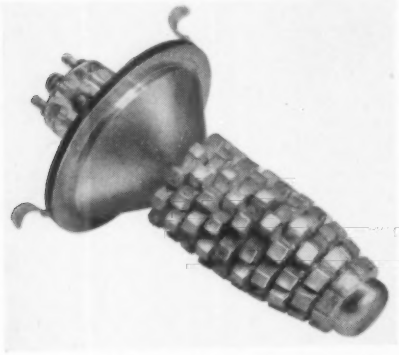
A wide-range a.f.c. system offers considerable advantages over the earlier combination of a narrow-range a.f.c. and the associated collection of preset adjustments. These include the elimination of a number of the presets, a reduction in the local-oscillator noise, and a great increase in the ability of the a.f.c. to deal with serious thermal or other similar drifts. It is necessary to weigh these advantages against the increased complexity of the a.f.c. and this point will be discussed after the circuit techniques have been described.

(To be continued)

References will appear at the end of the last instalment.

PHYSICAL SOCIETY'S EXHIBITION

THIS year's exhibition—the 38th—of scientific instruments and apparatus, which was held by the Physical Society from 8th to 13th April at Imperial College, London showed some change of character. In part this has been brought about by some reduction of the space available for the exhibition, for the consequent reduction in the number of exhibits was made much more in production apparatus than in research and development items. A welcome feature of this was the way in which it reduced the overlap with other exhibitions.



An example of the Ediswan Vapotron valve.

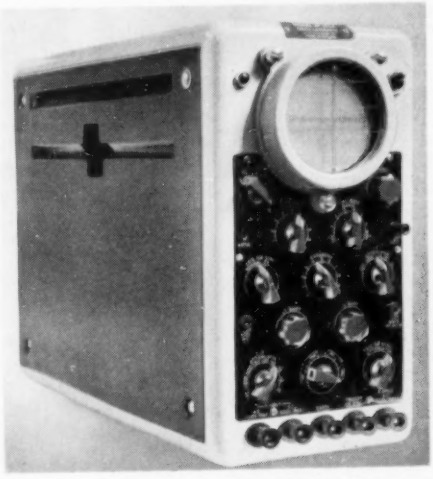
In the main, however, the change of character resulted from an alteration in the balance of physics itself. The post-war predominance of electronics is now passing and it is taking its proper place as a branch—albeit an important one—of physics. The trend is now away from an exclusive pre-occupation with the electron and what can be done with it and towards nuclear physics. Even here, however, in one form or another, the electron holds a definite place. Just as the electron is an essential part of the atom, so electronic apparatus seems essential to nuclear research. The electron, however, is now more a means to an end than an end in itself.

Viewed as a whole, this year's exhibition was more representative of physics than some previous ones have been, and the change is a very welcome one. Even so, the older branches of physics do not occupy the place they deserve.

It is, however, the field of electronics which is the main concern of *Wireless Engineer* in this exhibition and so this review of it will be confined to that field. Because of the smaller exhibition and because of the increase in the exhibits in other fields, the electronic section was a good deal smaller than for many years past.

Properly speaking, in its use as a noun, the word 'electronics' should be confined to valves and transistors, but it has been used above in its more popular sense as applying to anything which makes use of these devices. However, electronics in its true sense makes a good starting point for considering the exhibits and here an item of particular interest to the transmitting engineer is the Vapotron valve specially designed for use in the Vapodyne system of cooling, and shown by Ediswan. The anode is segmented externally in order to provide a large surface area and cooling is effected by water vapour. The valve heats the water to boiling point and utilizes the latent heat of evaporation. The advantages claimed for the system are that the heat dissipation is increased for a given size of valve, no pump is needed, the rate of water flow is small and insulation problems are eased.

At the other end of the scale, some new G.E.C. magnetrons are of interest. The VX3238 magnetron for 10,000 Mc/s is built in the usual miniature bulb and utilizes spatial harmonics of the resonator field; in this way operation can be obtained at the same voltage and magnetic field as a normal valve having 10–14 cavities.

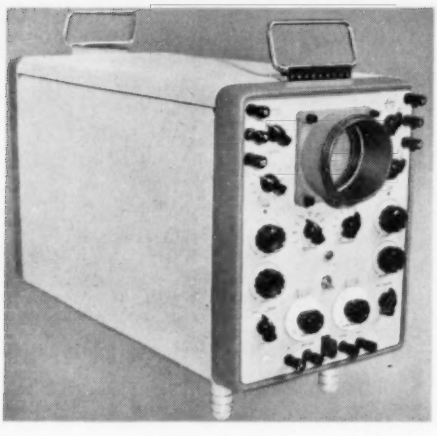


Cossor single-beam oscilloscope, type 1058.

G.E.C. magnetrons; (left) 3-*cm type* VX3238, (right) 10-*cm type* VX3209.

Only two cavities are used and an output of 250 mW may be obtained at 900 V and 7 mA with a field of 2,400 Oe. The output is directly into a waveguide and the frequency can be varied over a range of 400 Mc/s with a single tuning control.

The VX3209 is another magnetron for a somewhat lower frequency. It provides an output of 4–8 W over a tuning band of 8·8–11·6 cm and operates at 700–800 V, 30 mA with a magnetic field of 1,200 Oe. It is known as an interdigital magnetron, for the two anode elements



Furzehill oscilloscope intended for general-purpose use and in connection with servo systems.

each have six projecting fingers which interlace when the two are assembled facing each other. An external cavity resonator surrounds the anode structure and is adjustable for tuning. Operation is normally in the E_{010} mode, but the E_{020} mode can be used and with this an output of 2 W at 6·5 cm has been obtained.

As always in this exhibition, voltage stabilization forms a considerable part of the apparatus and for the higher voltages corona stabilizer tubes are now being made by several firms. One example is the Enarde (Nucleonic & Radiological Developments) tube; it is a gas-discharge tube operating in the corona régime. It is claimed to provide stabilization of some 0·5% for supplies of 400–1,300 V, the tube current being 6–120 μA at 400 V and 20–500 μA at 1,300 V. Better stabilization can be obtained by using tubes in cascade and higher voltages can be stabilized by connecting them in series.

Semi-Conductors

The transistor is now no longer a research novelty but, in some types, a production item. Mullard, for instance, showed a range of no less than two point-contact types and five junction types, of which the OC70 and OC71 are for use in production equipment and the OC10, OC11 and OC12 for experimental purposes. A Mullard demonstration of the hole-storage effect in germanium showed how an abnormally-high reverse current may occur immediately after forward-current flow. The output waveform of two diodes excited by a pulse was shown on an oscilloscope, the one having hole storage exhibited a marked overshoot at the end of the pulse with a subsequent slow decay, whereas the one without hole storage exhibited no overshoot in its output waveform.

Although the transistor is almost a normal piece of electronic equipment, some of its capabilities are by no means well known. G.E.C. had a demonstration of a transistor oscillator operating with an h.t. supply of 0·3 V at 7 μA, the 'battery' being two electrodes immersed in a jar of tap water! It is in this capability of functioning on a microscopic input power that the transistor is unique.

The application of germanium to power rectification was illustrated by part of the Standard Telephones & Cables exhibit. Such rectifiers have an exceedingly low voltage drop, 0·5 V mean only in the case of the R50A, which is rated for 100 mA d.c. output and 100 V r.m.s. maximum input. The zero voltage capacitance is 0·003 μF. This firm also showed germanium photocells which have characteristics similar to the vacuum and gas-filled cells. They are intended mainly for 'on-off' applications and the P50A has a sensitivity of 30 mA/lumen with a spectral response peaking at 1·7 micron and a weight of only 0·02 oz.

Oscilloscopes

The cathode-ray oscilloscope in one form or another is still ubiquitous for research in almost every sphere. New instruments differ from the older ones mainly in providing more facilities, an improved performance and smaller physical dimensions. The Cossor 1058 single-beam model is one example; the Y amplifier response is flat within 0·4 db from d.c. to 2·4 Mc/s and is down by just over 2 db at 3·5 Mc/s. A phantastron is used for the saw-tooth generator with a balanced amplifier providing a sweep expansion of up to five times. A 4-in tube is used and the signal sensitivity is 0·35–175 V/cm, variable by a stepped input attenuator and a continuous control.

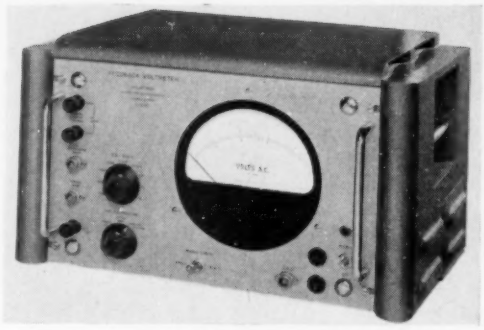
Mullard wideband oscilloscope, Mk. II, standing on its power-supply unit.

The complete instrument measures 8 in × 14½ in × 19½ in. A somewhat similar model (1052) with a double-beam tube is available.

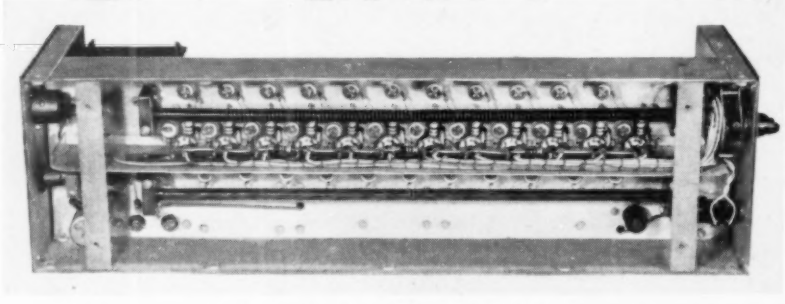
The difficulty of achieving accuracy in c.r. measurements is at last being realized and Cossor now make a voltage calibrator unit, Model 1433. It comprises a tapped transformer, attenuator and meter and enables known 50-c/s outputs of 3 mV to 100 V to be obtained with an accuracy of ± 5%.

A much more elaborate instrument was shown by Mullard, the E7581/2. The Y amplifier has a 6-db bandwidth of d.c. to 15 Mc/s and a sensitivity of 10 mV/cm; it is claimed to be able to handle a signal with a rise time of 0·03 μsec. A Miller time-base is used and is of the single-sweep type which may be triggered either by the signal or by an internal oscillator covering 2 c/s to 200 kc/s. A delay circuit is included by means of which the start of the sweep can be delayed by up to ten times the sweep duration.

It is built in two units with forced-air ventilation for the display unit; this unit has 42 valves, of which six are double types, and the power unit has 50. Needless to say, the power supplies are stabilized!



Solartron feedback voltmeter.



Underside of E.M.I. distributed amplifier.

It is always a difficult matter to secure sufficient output from a Y amplifier when a large bandwidth is needed. An amplifier shown by E.M.I. has this as one of its applications. It is a distributed amplifier having an output of 150 V p-p and a 6-db bandwidth of 30 c/s to 80 Mc/s and a gain of about 12 times.

An even more elaborate oscilloscope is the Southern Instruments E45 with a four-gun c.r. tube. There are four Y amplifiers, balanced and direct-coupled, with a gain of 800 times and a response to 100 kc/s. The time-base covers 1 c/s to 50 kc/s and there is a time-marker which generates pulses at four standard frequencies between 10 c/s and 1,000 c/s. The frequencies are generated by a multivibrator controlled by a tuning fork.

The complexity now shown by oscilloscopes is to be found in other instruments and one example is a voltmeter, VF252, shown by Solartron. It has a feedback circuit which eliminates the effects of changes in the forward/backward resistance ratios of the rectifiers. An absolute accuracy of ± 1% is claimed and the sensitivity is sufficient to permit useful readings to be obtained as low as 20 μV. The ranges are 1·5 mV to 15 V f.s.d. and the input impedance is 50 MΩ, with a frequency range of 10 c/s to 100 kc/s. All told, the instrument embodies 18 valves.

Another voltmeter, which can also be used as a general-purpose laboratory amplifier, is the Cintel signal-level meter. It has a response of 15 c/s to 100 kc/s within ± 0·5 db and a gain of 75 db, while the input attenuator has a range of 80 db. As a voltmeter, there are four ranges of 1 mV to 1 V r.m.s. f.s.d.

As a contrast to the elaborate multi-valve equipment, a simple unit with many useful applications was shown by J. Langham-Thompson. It is a cathode-follower stage, built as a small unit with self-contained batteries capable of giving 48 hours continuous operation. Input and output are by screened connectors. The maximum input is 10 V p-p with 100 MΩ and the output impedance is 4 kΩ.

The application of television in the laboratory was brought out in last year's exhibition and it is now

Cintel signal-level meter.

extended by a camera tube which is sensitive to ultra-violet light. In the equipment shown by E.M.I., a tube is used which is sensitive to wavelengths of 200 Å; it has an end window and mosaic plate of quartz. One application of the apparatus is to the magnification of images produced by the ultra-violet microscope.

An extremely ingenious way of heating rectifier filaments in high-voltage rectifier circuits, which is particularly advantageous when voltage-multiplier circuits are used, was shown by the Admiralty Signals and Radar Establishment. It is of application only when the supply is derived from an r.f. oscillator of, say, 50–300 kc/s. As shown in Fig. 1, the capacitance

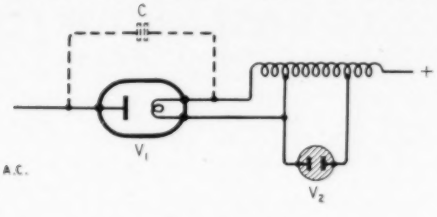
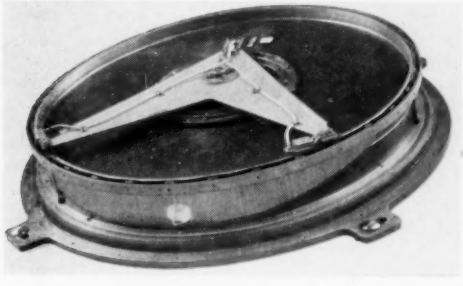


Fig. 1. *Circuit for obtaining filament-heating current for a high-voltage rectifier; C is the anode-filament capacitance.*



Precision sine-cosine potentiometer of 20-in diameter shown by Salford.

C of the rectifier V_1 passes a small alternating current. This is stepped up by an auto-transformer to a suitable value for the valve filament. If desired, a gas-stabilizer tube V_2 can also be connected to the transformer to stabilize the filament supply and so permit the d.c. output to be varied without affecting the filament current. The auto-transformer can be a simple compact device, since no high voltages appear between its parts.

Measuring Instruments

On the electrical side, there was as usual a display of precision components and instruments. Here it is interesting to see that, while the well-tried methods of construction are mostly retained, new methods are beginning to break in. Printed-circuit technique, for instance, is being used for decade resistance boxes by Tinsley. A printed foil method is used with copper-nickel foil, and the resistor values are adjusted to 1%. A three-decade box having 100-Ω, 10-Ω and 1-Ω steps is rated for a maximum current of 0·5 A.

Wayne Kerr capacitance and conductance bridge with direct-reading scales.

The decade principle is applied by Sullivan to an air-dielectric capacitor which comprises a 10-pF section continuously variable combined with two decade sections of 100 pF and 1,000 pF maximum capacitance each. These sections are obtained by variable capacitors of normal basic construction but having segmented vanes and a 'clicker' mechanism so that they move only in steps of 10 pF and 100 pF.

The use of sine-cosine potentiometers in predictors and similar apparatus is well known but few who are not intimately concerned with them realize the lengths to which it is necessary to go to secure the high accuracy required. The Salford types are based on the conventional

British Physical Laboratories Duo Meter; a multi-range test instrument with two meter movements and an unusual scale arrangement.

rotary wire-wound type but have a diameter of about 20 in! Nickel-chrome wire is used and wound on a card cut to the requisite shape and the contact wipers are of platinum ruthenium. The complete unit operates in an oil-filled case and the average accuracy of the law is claimed to be better than 0·07% with a life better than a million revolutions before resolution is affected.

The conventional form of dial and scale has been abandoned by some manufacturers in favour of an indicator of the counter type. One example is to be found in the Wayne Kerr B221 a.f. admittance bridge which is for the measurement of C and R at 10,000 radians/sec. It covers 0·1 pF to 0·1 μF. The indicator is direct-reading in picofarads to four figures and two decimal places and an interpolating drum for the third decimal place, so that there are seven figure-drums which give the value of any capacitance and conductance directly in figures. The range is extended by a multiplier of two decades.

In the case of pointer instruments, there would not appear to be much scope for any change in the styling but, in fact, quite a novel appearance has been obtained in the Duo Meter instrument produced by British Physical Laboratories. It is an instrument in the test-meter category having two independent movements with an unusual, but quite logical, arrangement of the 3½-in scales, as can be seen from the photograph.

The left-hand movement is for the voltmeter range and covers 1 V to 2·5 kV f.s.d. in eight d.c. ranges and 10 V to 1 kV in five a.c. ranges. There are also decibel scales of 0–62 db in five ranges. The right-hand movement is for current: 50 μA to 10 A f.s.d. in six ranges d.c. and 10 mA to 10 A in four ranges a.c. There are also three resistance ranges of 10 kΩ and 1 MΩ f.s.d.

Microwaves

In the microwave field, the measurement of power is still a matter of interest. Last year Elliott showed a method depending upon the rotation of a vane suspended in a waveguide section under the action of a mechanical couple exerted by the wave. This year's model for the 3·2-cm waveband, Type B228, can detect a minimum signal of 100 mW and has a peak power limit of 30 kW. The torque on the vane is measured by a calibrated quartz suspension and torsion head.

Wayne Kerr have introduced a model in which two vanes are used and mounted $\lambda/4$ apart on a spindle which is suspended in a length of vertical waveguide by a quartz fibre. The input and output guides are connected by means of E-plane corners. It is claimed that the accuracy of measurement is not greatly dependent on the matching to the load and that a standing-wave ratio of 0·9 causes an error of 0·5% only in the power indication.

An interesting filter for a coaxial line was shown by Decca Radar. It is formed from a number of radial cavities. A bandwidth of $\pm$ 4% is claimed for a mid-band frequency of 3,750 Mc/s and a total volume less than a 4-in cube. Attenuation away from the passband is of the order of 50 db.

NEW BOOKS

Ultra High Frequency Propagation

By Henry R. Reed, Ph.D., and Carl M. Russell, M.S. Pp. 562 + xiv. Chapman & Hall, Ltd., 37 Essex Street, London, W.C.2. Price 76s.

The title of this book is rather misleading, for what is usually understood by 'propagation' comprises a relatively small part of it. After an introductory chapter, the general aspects of propagation are discussed and then meteorological theory is treated. Chapter 4 is on aerials and radiation. Multipath propagation is dealt with in chapter 5 and then comes a comparison of v.h.f. and u.h.f. propagation, dipole-to-dipole, for ground-to-air and air-to-ground conditions. Aerial arrays are treated in the next two chapters, and certain complex forms, such as switched sector aerials, in a third. Chapter 10 covers air-to-air propagation and 11 with lobe modulation interference while the rest of the book treats more general matters.

The scope and character of the book might well be summed up by saying that its title should have been "The effect of propagation on the design and performance of telecommunication systems", with the accent on 'systems'. There is little or nothing about the actual apparatus at each end, the treatment is concerned with the communications link from the input to one aerial to the output of the other and it is concerned with everything that can affect this part of the chain.

Principles of Electronics

By L. T. Agger, B.E. Pp. 340. Macmillan & Co., Ltd., St. Martin's Street, London, W.C.2. Price 18s.

Starting with atomic structure, this book goes on to deal with electron dynamics, thermionic emission and conduction through gases. The diode and rectification, the triode and voltage amplification and multi-electrode valves and multi-stage amplifiers are then treated. Tuned amplifiers and power amplifiers follow and are succeeded by gas-filled valves, controlled rectification and inversion, oscillators, modulation and demodulation, cathode-ray tubes and photoelectricity.

The book is intended for the electrical student with an electrical and mathematical background corresponding to the Ordinary National Certificate in electrical engineering.

Circuit Theory of Electron Devices

By E. Milton Boone. Pp. 483 + vii. Chapman & Hall, Ltd., 37 Essex Street, London, W.C.2. Price 68s.

This is an American book "written primarily for college and university students". It covers ordinary valve types and their equivalent circuits, a.f. voltage and power amplifiers and gas-filled valves. Single and polyphase rectifiers are treated, r.f. voltage and power amplifiers with single- and double-tuned circuit couplings, as well as oscillators, modulation and demodulation. There is a final chapter on transistors.

Electronic Engineering Principles (2nd Edition)

By John D. Ryder, Ph.D. Pp. 505 + ix. Sir Isaac Pitman & Sons, Ltd., 39 Parker Street, London, W.C.2. Price 37s. 6d.

Of American origin, this book starts with the fundamental particles of electronics and goes on to deal with the movement of charged particles in fields and then to the cathode-ray tube. Emission comes in Chapter 4 and is followed by space-charge and then by diodes, triodes and multi-element valves. Small-signal amplifiers and feedback are then treated and, in Chapter 10, large-signal amplifiers. A jump is then made to gaseous conduction, gas diodes and gaseous control valves and circuits. The book concludes with photo-electric cells and solid-state electronics.

Théorie des Fonctions Aléatoires

By A. Blanc-Lapierre and Robert Fortet. Pp. 693 + xvi. Masson et Cie., 120 Boulevard Saint-Germain, Paris 6e, France. Price 6,500 fr.

After a physical introduction to random functions, there is a chapter on probability theory and the one on general random functions. Poisson processes and Markov processes and chains are treated. A chapter is given to the harmonic analysis of random functions and another to problems connected with it. Energy properties of, and stationary, random functions of second order are discussed as well as Laplacian random functions, the statistical theory of turbulence and the application of the functions to noise and telecommunications.

Manual for Plastic Welding. Vol. II. Polyethylene

By G. Haim and J. A. Neumann. Pp. 128 + xiv. Crosby Lockwood & Son, Ltd., 39 Thurloe Street, London, S.W.7. Price 30s.

This book describes the processes used, and the equipment necessary, for the welding of polyethylene. Although the processes are mainly ones appropriate to a commercial scale of operations, the use of hand tools is treated.

Higher Transcendental Functions

Compiled by the Staff of the Bateman Manuscript Project and based in part on notes left by Harry Bateman, late Professor of Mathematics, Theoretical Physics and Aeronautics at the Californian Institute of Technology. Vol. 1, pp. 302 + xxvi, price 52s. Vol. 2, pp. 396 + xvii, price 60s. McGraw-Hill Publishing Co., Ltd., 95 Farringdon Street, London, E.C.4.

The Magnetic Amplifier (2nd Edition)

By J. K. Reyner, A.C.G.I., B.Sc., D.I.C., M.I.E.E., M.Inst.R.E. Pp. 119. Rockliff Publishing Corporation, Ltd., 1 Dorset Buildings, Salisbury Square, Fleet Street, London, E.C.4. Price 15s.

This book gives a simple and substantially non-mathematical explanation of how magnetic amplifiers function. It provides an introduction to the subject.

Introduction à l'Électronique

By P. Grau. Pp. 212 + xi. Dunod, 92 rue Bonaparte, Paris 6e. Price 1,650 fr.

Cours Pratique de Télévision, Vol. 1. Amplificateurs M.F. et H.F. Directs à Large Bande.

By F. Juster. Pp. 127. Éditions Techniques et Professionnelles. 18 bis Villa Herran, Paris 16e. Price 490 fr.

Bessel Functions and Formulae

Compiled by W. G. Bickley. Pp. 11. Extracted from British Association Mathematical Tables, Vol. 10, Bessel Functions, Part 2, Functions of Positive Integer Order. Cambridge University Press, 200 Euston Road, London, N.W.1. Price 3s. 6d.

Radio Interference from Motor Vehicles. Comparison of British and German Measuring Equipment.

By A. H. Ball and S. F. Pearce. Technical Report M/T 123. Pp. 8. The British Electrical and Allied Industries Research Association, Thorncroft Manor, Dorking Road, Leatherhead, Surrey. Price 10s. 6d.

Wireless and Electrical Trader Year Book 1954 (25th Edition)

Pp. 296. Trader Publishing Co., Ltd., Dorset House, Stamford Street, London, S.E.1. Price 10s. 6d.

Les Filtres a Cristaux Piézoélectriques

By D. Indjoudjian and P. Andrieux. Pp. 177. Gauthier-Villars, 55 Quai des Grands-Augustins, Paris 6e, France. Price 3,300 fr.

Science in the U.S.A. 1953.

British Commonwealth Scientific Office, North America. Pp. 44 + iv. H.M. Stationery Office, York House, Kingsway, London, W.C.2. Price 2s.

Radio Control of Model Aircraft

By G. Sommerhoff. Pp. 164. Percival Marshall & Co., Ltd., 19-20 Noel Street, London, W.1. Price 9s. 6d.

Television Receiver Servicing. Vol. 1: Time-Base Circuits

By E. A. W. Spreadbury, M.Brit.I.R.E. Pp. 310. Trader Publishing Co., Ltd., distributed by Iliffe & Sons, Ltd., Dorset House, Stamford Street, London, S.E.1. Price 21s.

British Plastics Year Book 1954 (24th Edition)

Pp. 600. Iliffe & Sons, Ltd., Dorset House, Stamford Street, London, S.E.1. Price 30s.

The Magnitude of the Radio Interference in the Television Band from Ignition Systems of Motor Vehicles

By A. H. Ball and W. Nethercot. E.R.A. Bulletin U/T123. Pp. 11. The Electrical Research Association, Thorncroft Manor, Dorking Road, Leatherhead, Surrey. Price 6s.

Enregistrement des Sons

By J. Landrac. Pp. 232. Editions Eyrolles. 61 Boulevard Saint-Germain, Paris 5e, France. Price 1,900 fr.

The Radio Amateur's Handbook 1954 (31st Edition)

By the Headquarters Staff of the American Radio Relay League. Pp. 619. American Radio Relay League Inc., West Hartford 7, Connecticut, U.S.A. Price $3.

Dielectric Aerials

By D. G. Kiely, M.Sc., A.M.I.E.E., A.Inst.P. Pp. 132 + xii. Price 8s. 6d.

Magnetic Amplifiers

By George M. Ettinger. Pp. 88 + viii. Price 6s. 6d.

Microwave Spectroscopy

By M. W. P. Strandberg. Pp. 140 + vii. Price 9s. 6d.

The above three books are of the Methuen Monograph Series published by Methuen & Co., Ltd., 35 Essex Street, London, W.C.2.

The Year that made the Day

Pp. 79. The British Broadcasting Corporation, 35 Marylebone High Street, London, W.1. Price 6s.

An illustrated description of the planning and preparation for the B.B.C. broadcasts on Coronation Day.

Einführung in die Mikrowellen

By H. H. Klinger. Pp. 117 + viii. S. Hirzel Verlag, Zürich. Price 10.80 fr.

Problèmes et Exercices D'Électricité Générale et de Machines Électriques

By P. Janet, new edition revised by G. Nasse. Pp. 409 + vi. Gauthier-Villars, 55 Quai des Grands-Augustins, Paris 6e. Price 2,700 fr.

BRITISH STANDARDS:

Thermosetting Synthetic-Resin Bonded-Paper Insulating Sheets for use at Radio Frequencies

B.S. 2076 : 1954. Price 2s. 6d.

Characteristics and Performance of Apparatus for Measurement of Radio Interference

B.S. 727 : 1954. Price 4s.

Limits of Radio Interference

B.S. 800 : 1954. Price 4s.

Cotton Covered Round Copper Wires (Metric Units)

B.S. 2084 : 1954. Price 2s. 6d.

Determination of Power Factor and Permittivity of Insulating Materials by the Method of Hartshorn and Ward

B.S. 2067 : 1953. Price 4s.

Electrical Indicating Instruments

B.S. 89 : 1954. Price 7s. 6d.

Code of Practice on the Use of Electronic Valves (Parts 1 and 2)

C. P. 1005. Price 6s.

Glossary of Terms for the Electrical Characteristics of Radio Receivers

B.S. 2065 : 1954. Price 6s.

British Standards Institution, 2 Park Street, London, W.1.

NATIONAL BUREAU OF STANDARDS:

Simultaneous Linear Equations and the Determination of Eigenvalues

National Bureau of Standards Applied Mathematics Series 29. Pp. 126. Price $1.50.

This book contains 19 of the papers presented at a Symposium held at the National Bureau of Standards Institute for Numerical Analysis at Los Angeles on 23rd-25th August 1951.

Table of Coefficients for the Numerical Calculation of Laplace Transforms

National Bureau of Standards Applied Mathematics Series 30. Pp. 36. Price 25 cents.

Table of Natural Logarithms for Arguments between Zero and Five to Sixteen Decimal Places

National Bureau of Standards Applied Mathematics Series 31, (revision of MT10). Pp. 501. Price $3.25.

Magnetic Fields of Cylindrical Coils and Annular Coils

National Bureau of Standards Applied Mathematics Series 38. Pp. 29. Price 25 cents.

This booklet contains formulae for calculating the axial and radial components of the magnetic field at any point in space of certain coils. The results are expressed in terms of complete elliptic integrals or of Legendre functions.

Tables of 10^x

National Bureau of Standards Applied Mathematics Series 27. Pp. 543. Price $3.50.

Table I is for 10^x for $x = 0(\cdot 00001)\ 1\cdot 00000$ to 10D and Table II is a radix table of $10^{n.x}\ 10^{-p}$, $n = 1(1)999$, $p = 3, 6, 9, 12, 15$ to 15D. Antilogarithms (10^x) are given in an interpolable table to 10-decimal accuracy and in a basic radix table to 15-figure accuracy.

The National Bureau of Standards publications can be obtained from the Government Printing Office, Washington 25, D.C., U.S.A.

MEETINGS

Brit.I.R.E.

5th May. "Microwave Measuring Equipment", by P. M. Ratcliffe, to be held at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1, commencing at 6.30.

Television Society

14th May. "Receiver Design for 625-line Systems", by Dr. A. J. Biggs, to be held at the Cinematograph Exhibitors' Association, 164 Shaftesbury Avenue, London, W.C.2, at 7 o'clock.

Royal Society of Arts

5th May. "Colour Television", by Commander C. G. Mayer, O.B.E., to be held at John Adam Street, Adelphi, London, W.C.2, at 2.30.

STANDARD FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

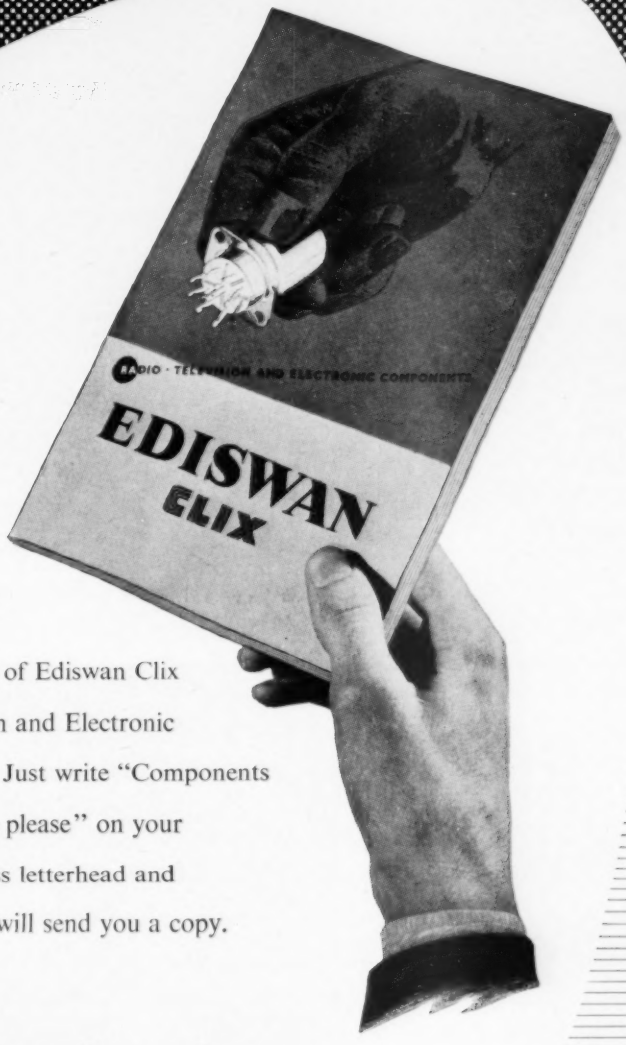
Values for March 1954

Date 1954 March	Frequency deviation from nominal: parts in 10^8		Lead of MSF impulses on GBR 1000 G.M.T. time signal in milliseconds
	MSF 60 kc/s 1429-1530 G.M.T.	Droitwich 200 kc/s 1030 G.M.T.	
1	−1·3	−1	+ 1·4
2	−1·3	0	+ 1·3
3	−1·3	+1	+ 1·3
4	−1·3	0	− 0·7
5	−1·2	0	− 1·0
6	−1·2	0	NM
7	−1·2	−1	NM
8	−1·3	−1	− 5·1
9	−1·3	0	− 6·4
10	−1·2	0	− 7·4
11	−1·2	−1	− 8·5
12	−1·2	−1	−10·2
13	−1·2	0	NM
14	−1·2	−1	NM
15	−1·1	0	−12·0
16	−1·1	+1	−13·5
17	−1·2	−1	−14·2
18	−1·2	+1	−18·2
19	−1·2	+1	−20·0
20	−1·2	−1	NM
21	−1·2	0	NM
22	−1·2	+1	−25·8
23	−1·2	+1	−27·7
24	−1·2	0	−30·1
25	−1·1	0	−31·9
26	−1·1	0	−33·3
27	−1·1	0	NM
28	−1·1	0	NM
29	−1·2	0	−39·8
30	−1·2	+1	−41·3
31	−1·0	+2	−42·6

The values are based on astronomical data available on 1st April 1954.

The transmitter employed for the 60-kc/s signal is sometimes required for another service.

NM — Not Measured.



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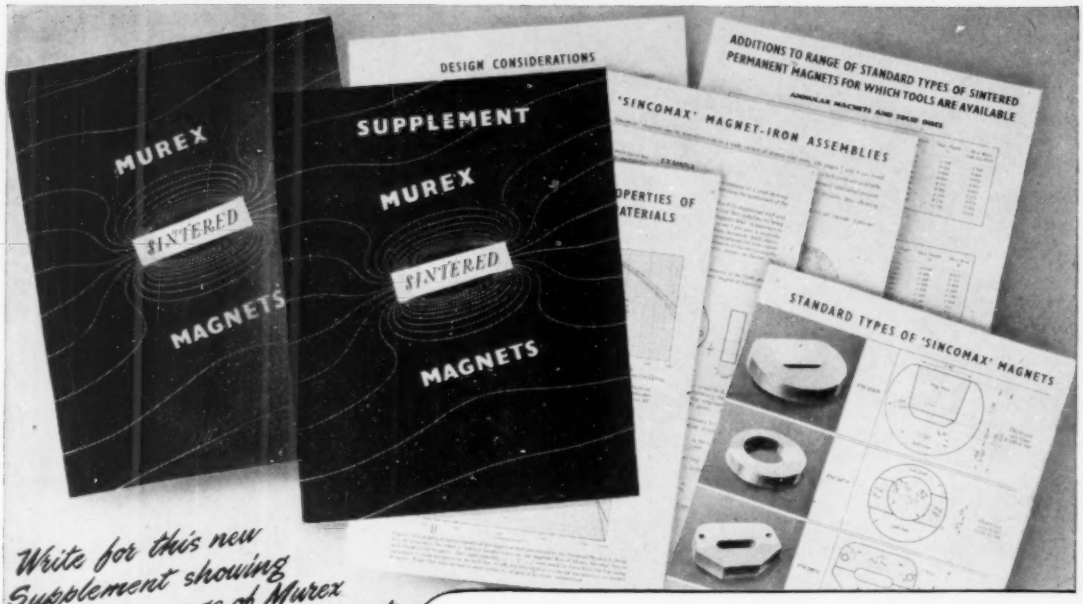

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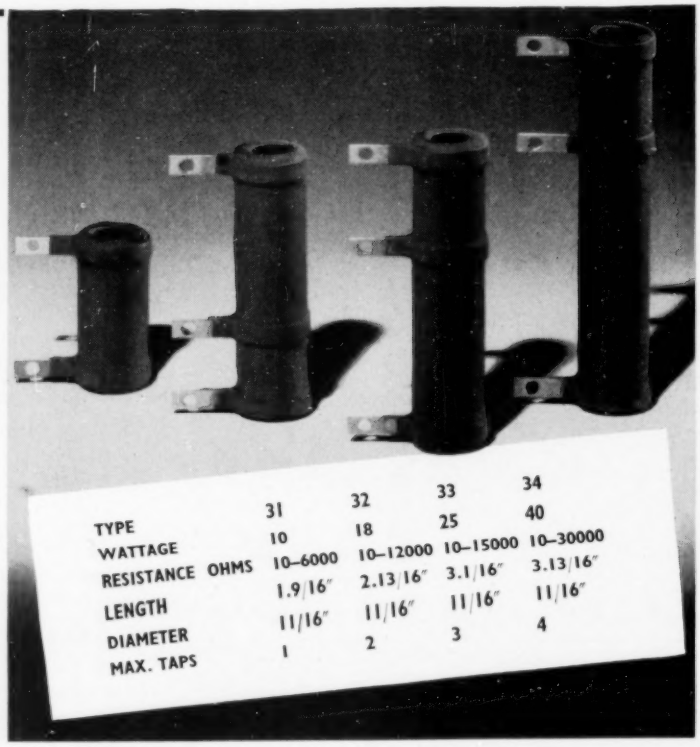
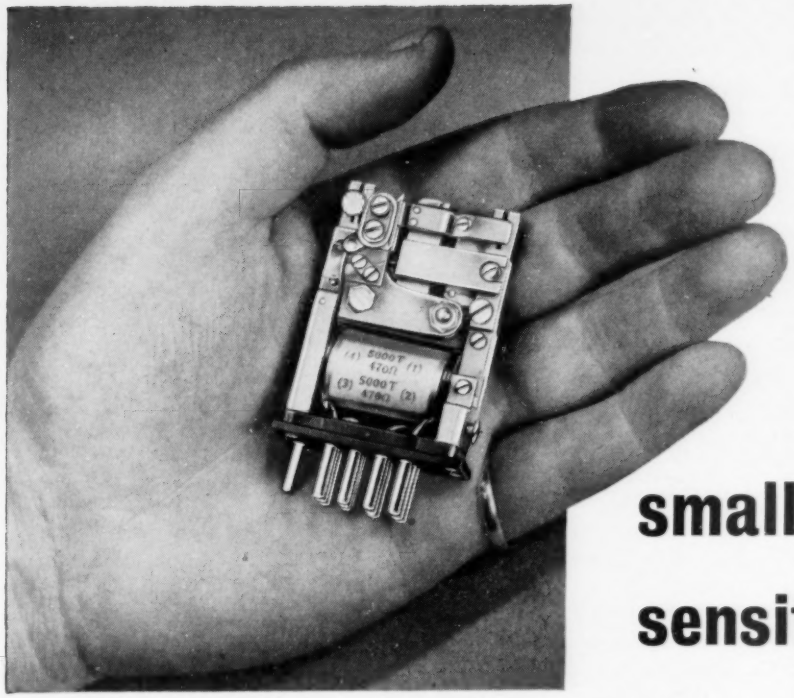


TYPE	31	32	33	34
WATTAGE	10	18	25	40
RESISTANCE OHMS	10–6000	10–12000	10–15000	10–30000
LENGTH	1.9/16″	2.13/16″	3.1/16″	3.13/16″
DIAMETER	11/16″	11/16″	11/16″	11/16″
MAX. TAPS	1	2	3	4



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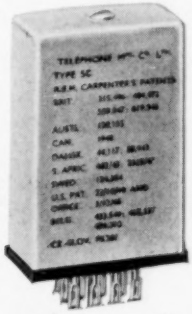

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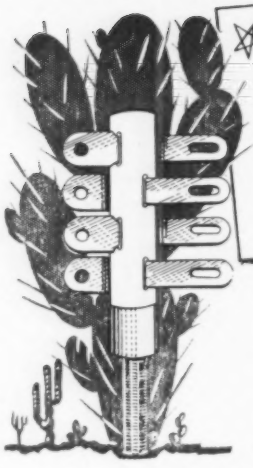

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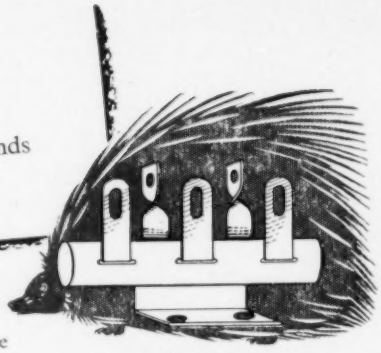

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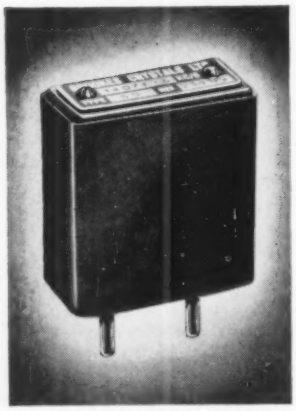

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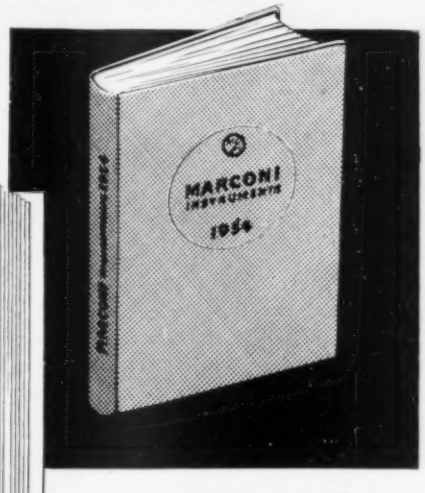

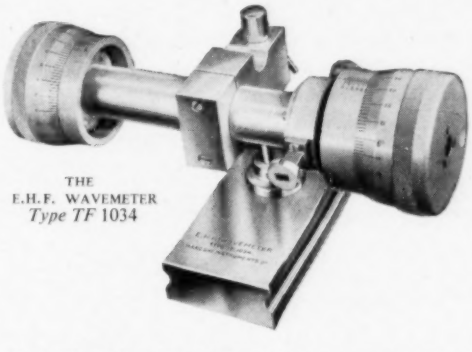

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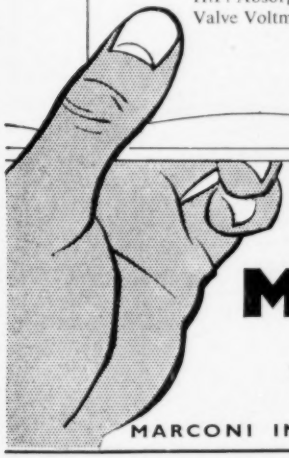

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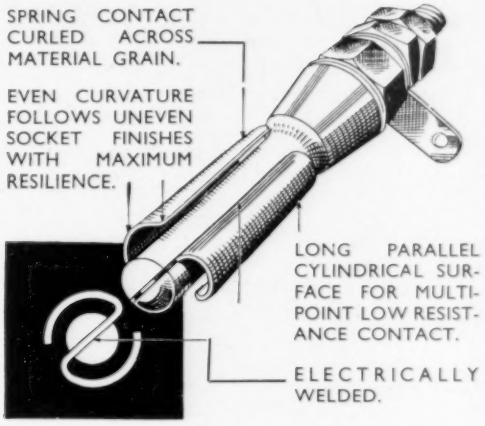
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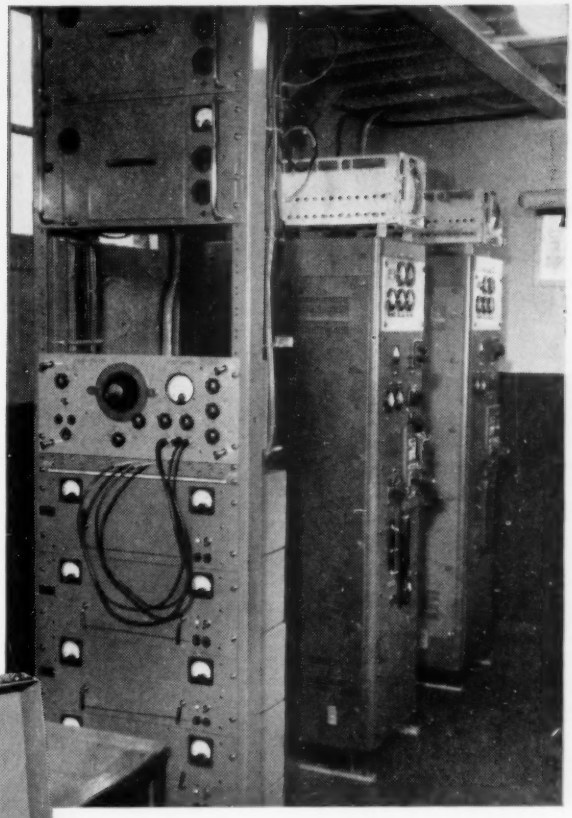

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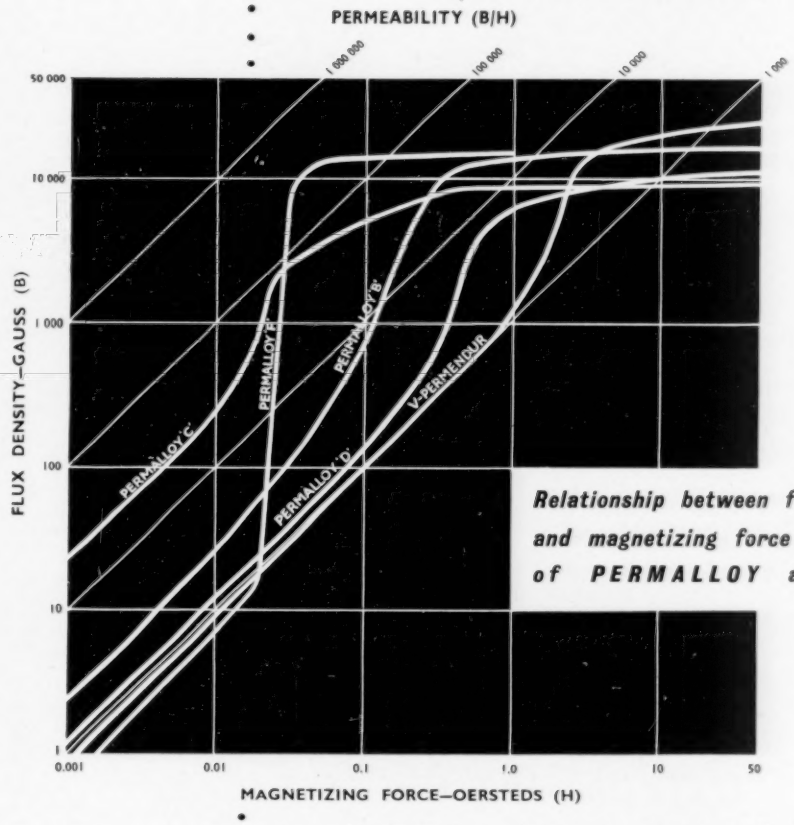

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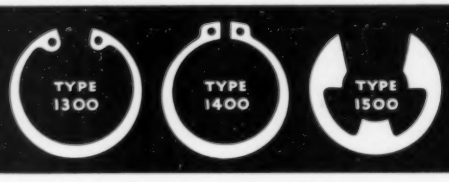

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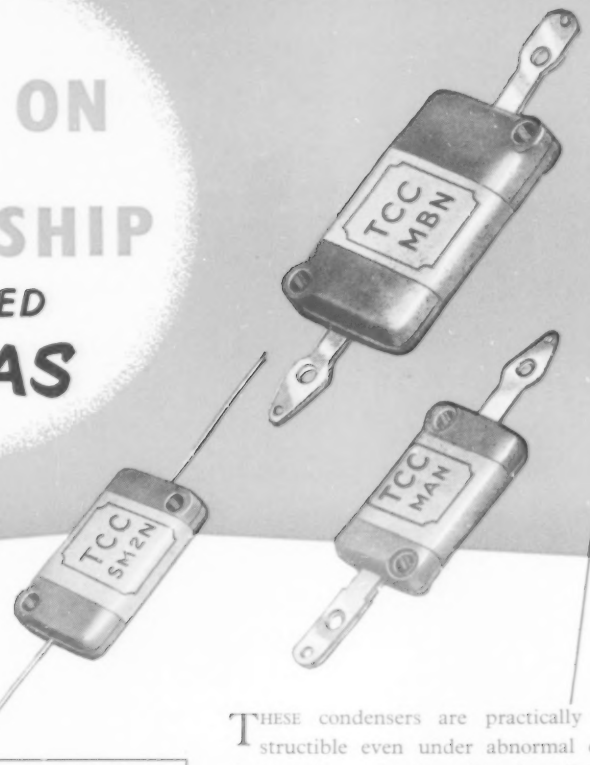

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